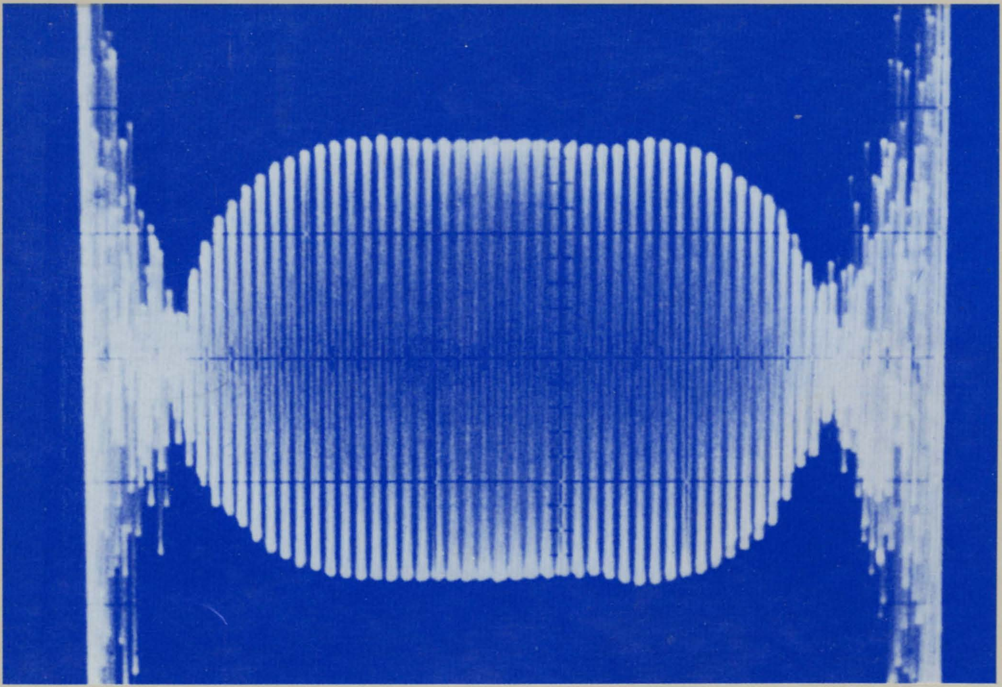
Elements of Electronic Instrumentation and Measurement

Second Edition



Joseph J. Carr

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Updated to Include Computer-Based
Instrumentation!

JOSEPH J. CARR, M.S.E.E.

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PREFACE TO THE 2nd EDITION

The first edition of *Elements of Electronic Instrumentation & Measurement* was quite successful—more so than we anticipated. Thus, when the publisher needed an updated book I was requested to revise my text for a 2nd edition. Since the first edition was published in 1978 there have been some changes in the instrumentation field. For one thing, digital electronics has become more important and less expensive. For another, digital electronic control is now widespread: the General Purpose Interface Bus (IEEE-488) makes automatic test equipment easier than before. Thus, a chapter on the IEEE-4888 GPIB has been added.

Although the book is intended for any course in electronic instrumentation—or for self-study—the principal focus is on the kinds of instrumentation required in broadcasting, audio, troubleshooting and maintenance, and in generalized testing. As in the first edition, a thoroughly practical element is maintained, but not at the expense of formal theory.

JOSEPH J. CARR, MSEE

PREFACE TO THE 1st EDITION

Even the least observant among us has to notice that we live in a world of measurement; in fact, it has been said that you really cannot understand a physical phenomenon until you can measure, i.e. quantify, it. A physicist friend of the author is fond of saying that physics is the science that measures reality.

Just as measurements are a tool for our understanding, instruments are a tool for measurements. Electronic instruments have proliferated to the extent that it is not possible for a person to claim a technical or scientific education without having some exposure to electronic instruments. People who never in the past had any use for instruments now find themselves heavily involved with an array of electronic devices. We find traditional analog and digital instruments widespread. The advent of the microprocessor adds near limitless capability; from small units inside of other instruments to larger number crunching systems.

In this book, you will study certain instrumentation techniques, both traditional and modern. You will be introduced to certain commercial products, and will come to understand *what* you are measuring. In certain cases, like the triggered sweep oscilloscope, skill in operating an instrument is so crucial to anyone in engineering, technology, or the sciences, that basic instructions are offered. You are, however, advised that it is best if you obtain as much practice as possible. If an electronics laboratory is available to you, then spend some time learning to operate all of the instruments . . . with special attention to the oscilloscope.

JOSEPH J. CARR

chapter 1

DC DEFLECTION INSTRUMENTS

1-1 OBJECTIVES

1. To learn the principles of operation of basic dc meter movements.
2. To learn methods for extending the current range of the basic dc meter.
3. To learn how to measure *voltage* by using a dc *current* meter.
4. To learn the operation of a volt-ohm-milliammeter.

1-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. Describe in your own words the *D'Arsonval* meter movement.
2. What is the principal difference between the *D'Arsonval* and *taut-band* meter movements?
3. How may a dc current meter be used to measure an ac voltage?
4. A ____________ resistor is used to make a dc current meter into a dc voltmeter.
5. A ____________ resistor is used to increase the range of a dc current meter.

1-3 THE BASIC DC METER

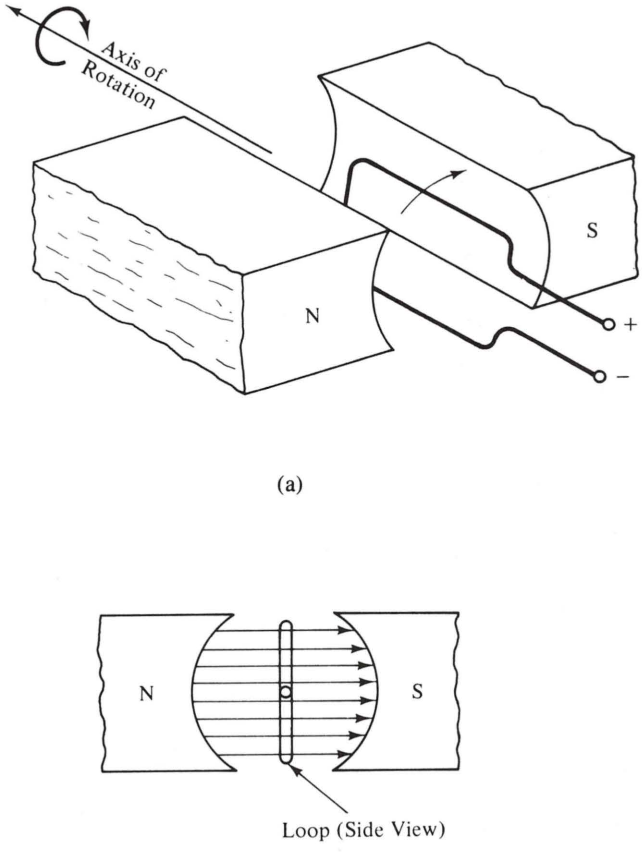
For many decades the basic dc meter movement has been used as the readout device in most electronic instruments, even in many instruments that measured ac. Digital readout devices were too expensive for general application until the early 1970s, so most instruments relied on analog meter movements to provide the readout function.

The two most common dc meter movement configurations are the *D'Arsonval* and *taut-band* designs. Both are examples of a *permanent magnet moving* *coil* (PMMC) galvanometer, and work on the same fundamental electrical principle as the dc motor.

A simplified view of the PMMC galvanometer is shown in Figs. 1-1(a) and 1-1(b): a movable coil of wire is mounted in the magnetic field between the poles of a permanent magnet.

A current flowing in a wire generates a magnetic field. The *polarity* of the field is determined by the *direction* of the current flow, while the *strength* of the field is determined by the *magnitude* of the current.

The coil in a PMMC galvanometer is mounted so that it can rotate in the space between the magnet poles. A current in the coil creates a magnetic field



(a)

(b)

Figure 1-1. (a) Construction of a D'Arsonval meter movement. This meter uses a loop of wire rotating in a magnetic field, much like an electrical motor. (b) Side view of the D'Arsonval meter movement.

that either *aids* or *opposes* the field of the nearby poles. A current flow in one direction causes a clockwise rotation, whereas a current flow in the opposite direction causes a counterclockwise rotation. The amount of rotational position change is proportional to the magnitude of the current.

1-4 THE D'ARSONVAL MOVEMENT

The *D'Arsonval* meter movement is shown in Fig. 1-2. A side view of the meter, without the permanent magnet, is shown in Fig. 1-2(a), while a frontal view *with* the magnet is shown in Fig. 1-2(b).

The coil in Fig. 1-2(a) is shown wound on an *armature* or *bobbin,* which is mounted on a pair of jewel bearings to reduce friction. This assembly is shown frontally in Fig. 1-2(b).

When a current flows in the coil, the armature assembly deflects *clockwise* [as viewed in Fig. 1-2(b)] an amount proportional to the strength of the current. The amount of deflection can be marked off in units of *current* on the dial scale. The coiled *pivot spring* serves to dampen the pointer movement, and to return the pointer to the zero position when the current flow in the coil ceases.

The travel of the pointer is limited by high and low end mechanical stops just beyond the *zero* and *full-scale* limits printed on the dial scale. The wires to the coil are given just enough slack that they will not be stretched anywhere in the pointer's normal range of travel.

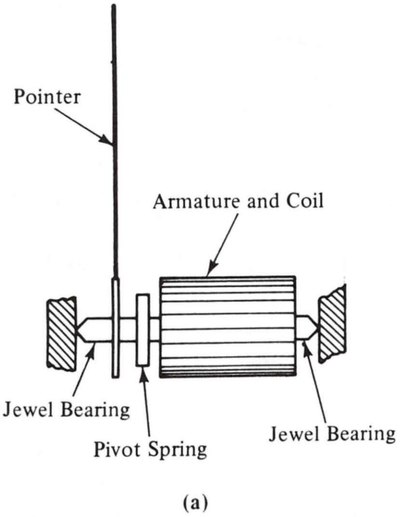


Figure 1-2. (a) Construction of the coil and pointer assembly in a D'Arsonval meter movement.

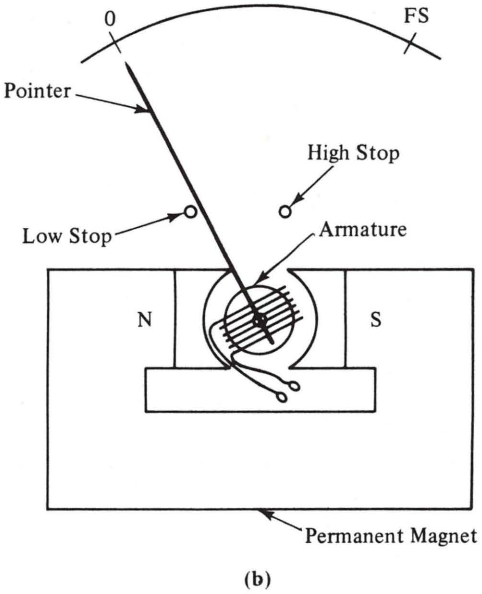


Figure 1-2. ***(cont.)*** (b) Usual configuration for a D'Arsonval meter movement.

1-5 THE TAUT-BAND MOVEMENT

The taut-band meter movement is essentially the same as the D'Arsonval, except for the manner in which the armature coil is mounted (see Fig. 1-3). In the taut-band movement the armature is *suspended* from fixed supports on a stretched, i.e., taut, rubber band. The band is twisted as the armature rotates, so no restoring force from a pivot spring is needed.

There are two principal advantages to the taut-band meter movement over the older D'Arsonval design: greater *sensitivity* and more *durability*.

Older D'Arsonval movements are rarely found with full-scale sensitivities of less than 50 microamperes, whereas there are taut-band models available that boast full-scale sensitivity of only 2 microamperes.

D'Arsonval movements are more easily damaged than taut-band types because of their jewel bearing construction. Even a short fall to the floor or a table top is sufficient in many cases to break the bearings. The rubber band in a taut-band meter can snap, but this occurs far less frequently than does bearing damage.

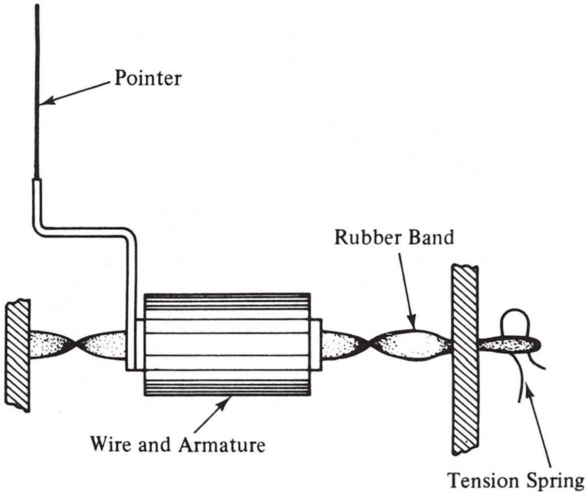
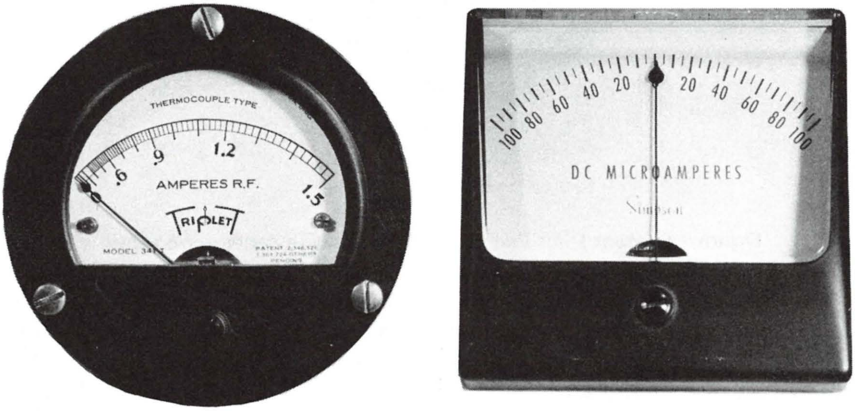


Figure 1-3. Construction of taut-band meter movement.

1-6 TYPES OF METERS

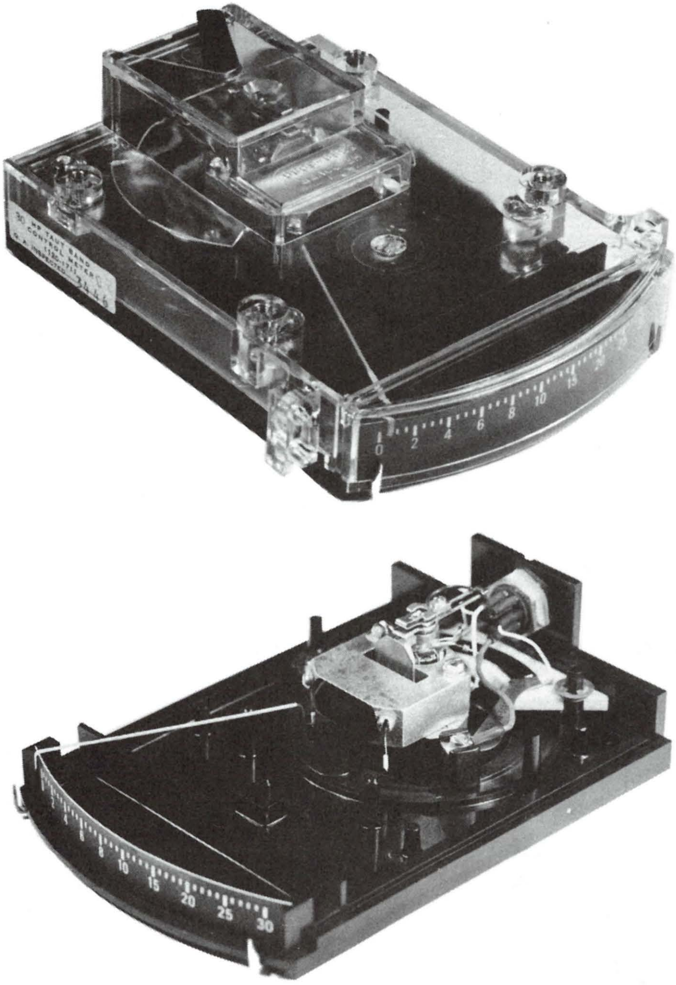
The two basic meter movements are available in an almost endless variety of sizes and shapes, but we can classify meter types by the form of scale that is used. Figure 1-4 shows three different configurations.



(a)

(b)

Figure 1-4. (a) Thermocouple RF ammeter. (b) Zero-center dc microammeter.



(c)

Figure 1-4. ***(cont.)*** (c) Two views of an edgewise meter movement.

The meter in Fig. 1-4(a) is the type normally encountered. The zero-current mark is on the left-hand side of the scale, and the pointer deflects to the right when a current flows. The input terminals for this type of meter are marked "+" and "−", and must not be reversed, or the pointer may be damaged when it attempts to deflect further to the left.

Figure 1-4(b) shows an example of the *zero-center galvanometer;* the zero-current mark on this type of meter is in the exact center of the scale. Current flowing in one direction will deflect the pointer to the left, while current flowing in the opposite direction will deflect the pointer to the right. When the polarity markings on the input terminals are observed, a positive current deflects the pointer to the right and a negative current deflects it to the left.

The type of meter movement shown in Fig. 1-4(c) is an *edgewise* panel meter. Its design allows conservation of *area* on the front panels of equipment in which the movement is used. Of course, this saving is at the expense of increased depth. The high and low limits are sensed by external circuitry and are set by the tabs on the meter face. Edgewise meters are available in both left-zero and center-zero models.

1-7 USING BASIC DC METERS

There are three basic rules for using dc *current* meters:

1. Connect the meter in *series* with the load or circuit in which the current is being measured.
2. Use the meter with a full-scale rating greater than the maximum expected current.
3. Use a meter that has an internal resistance that is *low* compared with the resistance of the circuit in which it is being used.

A current meter is *always* connected in *series* with the load. There are never any exceptions. *Failure to observe this rule may result in permanent damage to the instrument.* In cases where the meter is used to measure current in a branch of a larger circuit, then the meter is to be connected in series with that branch.

Figure 1-5 shows the correct connection for the 0 to 1 mA dc current meter. It is necessary to break the circuit, i.e., the wire to resistor $R1$, and connect the meter in series.

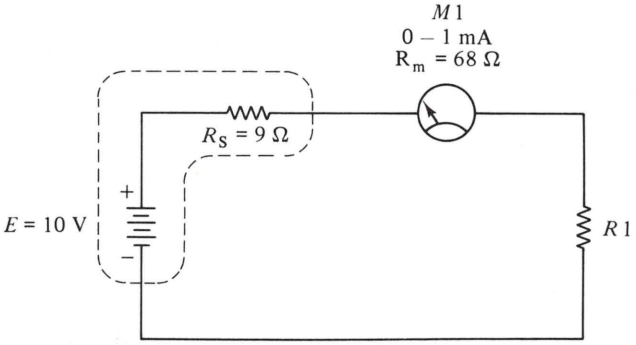


Figure 1-5. Circuit with 0–1 mA milliammeter.

The current flowing in meter $M1$ is given by

$$I_{m1} = \frac{E}{R_s + R_m + R1} \tag{1-1}$$

where I_m = the meter current in amperes (A)
E = the open-circuit voltage (V)
R_s = the power supply *source resistance* in ohms (Ω)
R_m = the internal resistance of the meter in ohms (Ω)
$R1$ = the load resistance in ohms (Ω)

In most cases R_s and R_m are very small with respect to $R1$, so are considered negligible. If that is true, then Eq. (1-1) reduces to Ohm's law

$$I_{m1} = \frac{E}{R1} \tag{1-2}$$

Example 1-1

Find the current flowing in $M1$ if $R1$ is 15 kΩ.

SOLUTION

[Since $R1$ is very much greater than R_s or R_m, use Eq. (1-2).]

$$\begin{aligned} I_{m1} &= E/R1 \qquad (1\text{-}2) \\ I_{m1} &= (10\ \text{V})/(1.5 \times 10^4\ \Omega) \\ I_{m1} &= 6.7 \times 10^{-4}\ \text{A} = \mathbf{0.67\ mA} \end{aligned}$$

The current in Example 1-1 is less than the full-scale rating of the meter—i.e., 1 mA—so our circuit obeys rule 2.

Example 1-2

Find the current flowing in $M1$ if it were *incorrectly* connected in parallel with $R1$ instead of in series as is proper.

SOLUTION

[Use Eq. (1-1) because $R1$ is very much greater than R_s and R_m so its parallel effect is negligible; delete $R1$ from the equation.]

$$I_{m1} = \frac{E}{R_s + R_m} \qquad (1\text{-}1)$$

$$I_{m1} = \frac{10\ \text{V}}{(9 + 68)\ \Omega} = 0.13\ \text{A} = \mathbf{130\ mA}$$

Connecting $M1$ across the load results in a current that is 130 times too high. This will destroy the instrument.

Rule 3 can become troublesome in low voltage, low resistance circuits, where the meter's coil resistance (R_m) is a significant portion of the load resistance. Neglecting R_s, let us see what would happen if $R1$ were reduced to 150 ohms, and E were only 100 mV. By Ohm's law, we would expect a current of

$$I = E/R1$$

$$I = (0.1\ \text{V})/(150\ \Omega)$$

$$I = 4.6 \times 10^{-4} = \mathbf{0.46\ mA}$$

No measuring or testing instrument should be used if it significantly alters the circuit performance, and in this case a 30 percent meter reading error was created by an unwise selection of the meter. In both cases we expected to find a reading of 0.67 mA, but in the latter case the meter reading was 0.46 mA.

1-8 OBTAINING HIGHER CURRENT SCALES

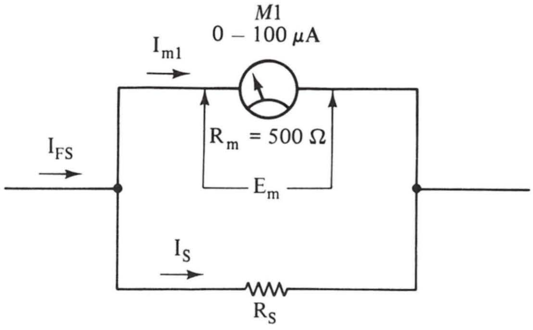
The basic dc meter movement has a single current scale (e.g., 0–1 mA, 0–200 μA, etc.), but can be used to measure larger currents if a *shunt resistor* in parallel with the meter is used (see Fig. 1-6).

The wire diameter required for the coil in high current meter movements is excessive and proves impractical, so it has become the standard practice to use a low-range meter movement and the shunt resistor.

In some cases the shunt is inside of the meter case, whereas in others it is located outside of the case, usually mounted to the meter's input terminals. An example using the shunt resistor is given in Fig. 1-6. The total full-scale current I_{fs} is given by

$$I_{fs} = I_{m1} + I_s \qquad (1\text{-}3)$$

where I_{fs} = the total current
I_{m1} = the current flowing in the meter
I_s = the current flowing in the shunt resistor



$$I_{m1} = I_{FS}\left[\frac{R_S}{R_m + R_S}\right]$$

Figure 1-6. Principle of the shunt resistor to increase current range.

Example 1-3

The circuit of Fig. 1-6 is used to measure 1 mA full scale. How much current (at full scale) flows in (a) $M1$ and (b) R_s?

SOLUTION

Solve Eq. (1-3) for I_s.

$$I_{fs} = I_{m1} + I_s \tag{1-3}$$

Therefore

$$\begin{aligned} I_s &= I_{fs} - I_{m1} \\ I_s &= [1\text{ mA} \times (10^3\ \mu\text{A}/1\text{ mA})] - 10^2\ \mu\text{A} \\ I_s &= (10^3 - 10^2)\ \mu\text{A} = \mathbf{900\ \mu A} \end{aligned}$$

There are two methods for calculating the value of R_2 if R_m, I_{m1}, and I_{fs} are known: *Ohm's law* and the *current divider equation.*

If we know the full-scale rating of the meter movement [in Example 1-3 it is 100 μA (Fig. 1-6)], and the meter coil resistance R_m, then we know by Ohm's law that the voltage drop across $M1$ at full scale is $I_{m1} \times R_m$, or in this case

$$\begin{aligned} E_{m1} &= (10^{-4}\text{ A})(5 \times 10^{12}\ \Omega) \\ E_{m1} &= 0.05\text{ V} \end{aligned}$$

Since R_s is connected in parallel with $M1$, the *same* voltage is across R_s. If we calculate I_s [Eq. (1-3)], and use Ohm's law, we can find the *value* required of R_s. In the case of Example 1-3, in which 900 μA flows in R_s,

$$R_s = E_{m1}/I_s$$
$$R_s = (0.05\ \text{V})/(9 \times 10^{-4}\ \text{A})$$
$$R_s = 55.56\ \Omega$$

The step-by-step procedure for using Ohm's law to determine the value of the shunt resistor is

1. Determine I_{fs}, I_{m1}, and R_m.
2. Calculate I_s, using Eq. (1-3).
3. Calculate E_{m1}, using Ohm's law.
4. Use Ohm's law again to calculate R_s from E_{m1} and I_s.

Or, in equation form:

$$R_s = \frac{I_{m1} R_m}{(I_{fs} - I_{m1})} \tag{1-4}$$

The current divider method takes advantage of Kirchoff's current law, which tells us that the current in R_s is a fraction of I_{fs}.

$$I_s = I_{fs} \times \left[\frac{R_m}{R_m + R_s} \right] \tag{1-5}$$

Example 1-4

A 0 to 50 μA dc meter movement, with a coil resistance R_m of 1250 ohms, is used with a shunt resistor to measure a full-scale current of 500 μA. Calculate R_s.

SOLUTION

Solve Eq. (1-5) for R_s.

$$I_s = (I_{fs})\,[R_m/(R_m + R_s)] \tag{1-5}$$
$$(500\ \mu\text{A} - 50\ \mu\text{A}) = (500\ \mu\text{A})\ [1250\ \Omega/(1250 + R_s)\Omega]$$
$$\frac{450}{(500)(1250)} = \frac{1}{1250 + R_s}$$

By algebraically rearranging the expression, we obtain

$$R_s = [(500)(1250)(450)]\ \Omega - (1250)\Omega$$
$$R_s = (1389 - 1250)\ \Omega = \mathbf{139\ \Omega}$$

Although each method for determining the value of the shunt resistor has its own proponents, either method will yield the same answer. The reader is encouraged to work Example 1-3 using the Ohm's law method.

1-9 VOLTAGE MEASUREMENTS FROM DC CURRENT METERS

Voltage can be measured on a dc current meter if a *multiplier resistor* is connected in series with the meter movement. This type of circuit is shown in Fig. 1-7(a), with the equivalent circuit shown in Fig. 1-7(b).

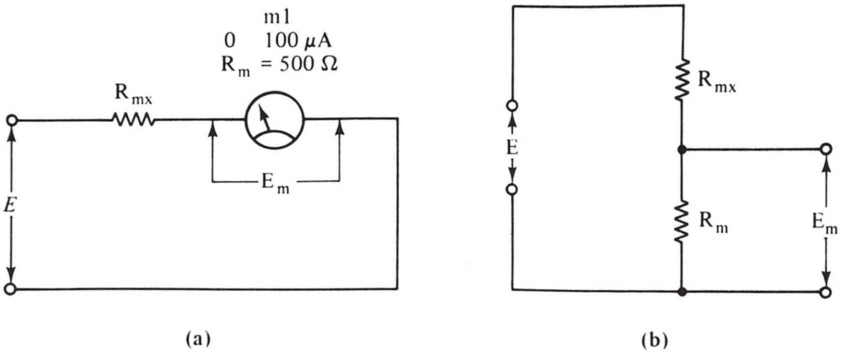


Figure 1-7. (a) Use of multiplier resistor to make an ammeter read in volts. (b) Equivalent circuit.

Input voltage E will set up a current $I1$ in the circuit of Fig. 1-7(a) that has a value of

$$I1 = E/(R_m + R_{mx}) \tag{1-6}$$

where $I1$ = the current in amperes (A)
E = the applied potential in volts (V)
R_m = the meter resistance in ohms (Ω)
R_{mx} = the multiplier resistance in ohms (Ω)

We may calculate the multiplier needed for any given full-scale voltage E by setting $I1$ in Eq. (1-6) equal to the full-scale rating of $M1$ and then solving Eq. (1-6) for R_{mx}.

Example 1-5(a)

Calculate the value of the multiplier resistor R_{mx} in Fig. 1-7(a) for a full-scale potential E of 10 volts.

SOLUTION

Solve Eq. (1-6) for R_{mx}.

$$I1 = E/(R_m + R_{mx}) \tag{1-6}$$
$$10^{-4}\ \text{A} = (10\ \text{V})/(500 + R_{mx})\ \Omega$$
$$10^{-5} = 1/(500 + R_{mx})\ \Omega$$

By rearranging the expression we obtain

$$R_{mx} + 500\ \Omega = 10^5\ \Omega$$
$$R_{mx} = 9.95 \times 10^4\ \Omega = \mathbf{99{,}500\ \Omega}$$

We may also use the *voltage divider equation* to calculate R_{mx}. We know that the full-scale voltage drop across $M1$ will be equal to $I1 \times R_m$, or, in the case of Example 1-5,

$$E_m = I1 \times R_m$$
$$E_m = (10^{-4}\ \text{A})(5 \times 10^2\ \Omega)$$
$$E_m = \mathbf{0.050\ volt}$$

So, by the voltage divider equation, we find a solution for Example 1-5.

Example 1-5(b)

Work the problem in Example 1-5(a) using the voltage divider method.

SOLUTION

$$E_m = \frac{E \times R_m}{R_m + R_{mx}} \tag{1-7}$$
$$(0.05) = (10)(500)/(500 + R_{mx})$$
$$\frac{(0.05)}{(10)(500)} = \frac{1}{500 + R_{mx}}$$

By rearranging the expression, we obtain

$$R_{mx} + 500 = (10)(500)/0.05)$$
$$R_{mx} = 100{,}000 - 500 = \mathbf{99{,}500\ \Omega}$$

1-10 VOLTMETER SENSITIVITY (Φ)

The *sensitivity* (Φ) of a voltmeter is given in terms of ohms per volts (Ω/V), and is dependent upon the full-scale current range of the basic dc meter movement used to make the voltmeter. The sensitivity can be found by taking the reciprocal of the meter's full-scale current rating.

$$\Phi = \frac{1}{I_{\mathrm{fs}}} \tag{1-8}$$

Proof: $\quad \Phi = 1/I_{\mathrm{fs}}, \quad \text{but} \quad I_{\mathrm{fs}} = E_{\mathrm{m}}/R_{\mathrm{m}}$

so $\quad \Phi = 1/(E_{\mathrm{m}}/R_{\mathrm{m}}) = R_{\mathrm{m}}/E_{\mathrm{m}} = \Omega/\mathrm{V}$

Table 1-1 gives the sensitivity ratings for voltmeters made by using various values of full-scale meter current.

Table 1-1

Full-scale Meter Current	*Sensitivity* (Φ)
1 mA	1000 Ω/V
100 μA	10 kΩ/V
50 μA	20 kΩ/V
20 μA	50 kΩ/V
10 μA	100 kΩ/V

1-11 VOLTMETER IMPEDANCE

The *impedance* of a voltmeter is the product of its full-scale voltage and the *sensitivity:*

$$Z = E_{\mathrm{fs}} \times \Phi$$

where Z = the input resistance in ohms (Ω)
E_{fs} = the full-scale rating of the meter in volts (V)
Φ = the sensitivity in ohms/volt (Ω/V)

Example 1-6

Calculate the input impedance of a voltmeter such as Fig. 1-7(a) on the 0-to-500-volt dc scale, if the sensitivity is 10 kΩ/V.

SOLUTION

$$Z = E_{fs} \times \Phi \tag{1-9}$$

$$Z = (500\ \text{V})(10^4\ \Omega/\text{V})$$

$$Z = 5 \times 10^6\ \Omega = \mathbf{5\ meg\Omega}$$

1-12 USING VOLTMETERS

Voltage is defined as the *difference* in electrical potential between two points in a circuit. Therefore, to measure voltage, connect the voltmeter *across* the load, i.e., in parallel with the load or circuit branch that you want to measure. Three basic rules apply:

1. Connect the voltmeter in parallel with the load (see Fig. 1-8).
2. Select a voltmeter that has a full-scale range that is greater than the highest expected voltage.
3. Make sure that the voltmeter input impedance (Sec. 1-11) is *very high* compared with the circuit resistance.

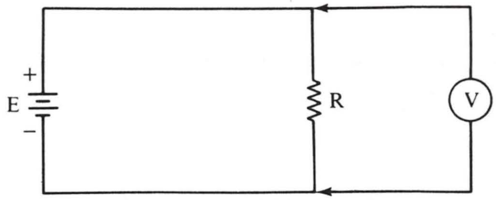


Figure 1-8. A voltmeter is always connected in parallel with the load.

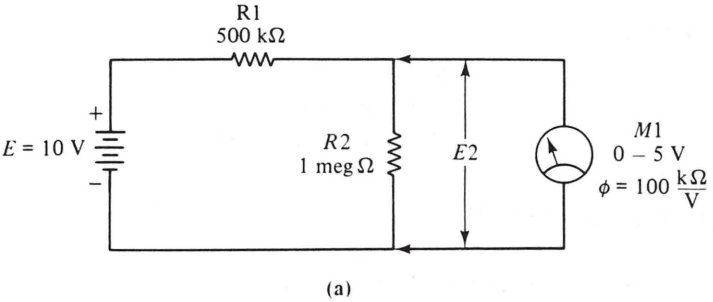
The voltmeter is a high impedance device, so any attempt to connect it in series will greatly increase circuit resistance, causing improper operation.

Problems of circuit loading occur if rule 3 is not obeyed. Consider Fig. 1-9(a). From the voltage divider equation we know that $E2$ should be

$$E2 = \frac{ER2}{R1 + R2}$$

$$E2 = \frac{(10\ \text{V})(1\ \text{meg}\Omega)}{1\ \text{meg}\Omega + 0.5\ \text{meg}\Omega}$$

$$E2 = (10\ \text{V})(1/1.5) = \mathbf{6.67}$$



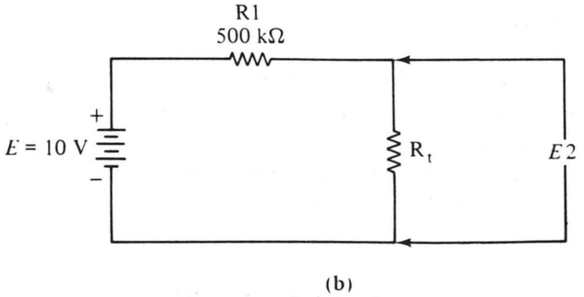


Figure 1-9. (a) Voltmeter impedance can cause errors due to circuit loading. (b) Equivalent circuit.

But at 0 to 5 volt full scale, $M1$ has an impedance of only 500 kΩ. When $M1$ is connected across $R2$, then, it is equivalent to shunting a 500-kΩ resistor across $R2$; the equivalent circuit is shown in Fig. 1-9(b). With R_m in parallel with $R2$, the parallel combination has a resistance of

$$R_t = \frac{R2(R_m)}{R2 + R_m}$$

$$R_t = \frac{(1\ \text{meg}\Omega)(0.5\ \text{meg}\Omega)}{(1\ \text{meg}\Omega + 0.5\ \text{meg}\Omega)}$$

$$R_t = 0.33\ \text{meg}\Omega$$

Again using the voltage divider equation, but this time with R_t substituted for $R2$, we find that voltage $E2$, as indicated on the meter, is

$$E2 = (E)(R_t)/(R1 + R_t)$$
$$E2 = (10\text{ V})(0.33)/(1 + 0.33)$$
$$E2 = (10\text{ V})(0.33/1.33) = \mathbf{3.75\ V}$$

A 44 percent error resulted from using a meter with an impedance that was too low for the circuit impedance, i.e., $R2$. The general rule is to select a meter with an input impedance that is not less than ten times greater than the circuit impedance.

1-13 OHMMETERS

Resistance can be measured by using a dc current meter and a calibrated voltage source; an example is shown in Fig. 1-10.

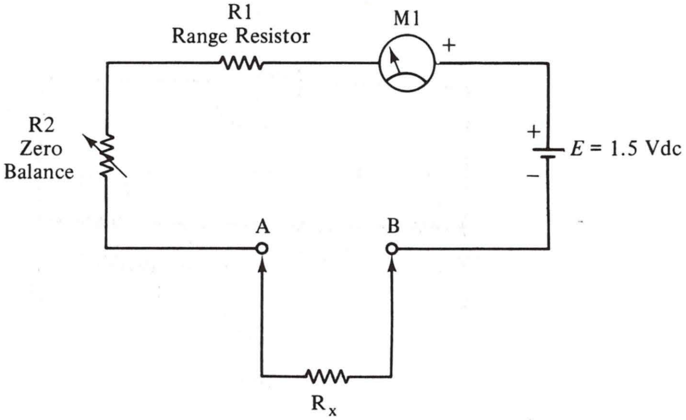


Figure 1-10. Simple ohmmeter.

Resistor $R1$ sets the range of the ohmmeter, while $R2$ is used as a *zero balance* control. $R2$ is adjusted for full-scale deflection when points A and B are shorted together, i.e., $R_x = 0$. This point is labeled "0-ohms" on the meter scale [see Fig. 1-11(a)]. If an unknown resistance is connected across terminals A-B, then the current through the meter is reduced proportionately to the value of the resistance. Figures 1-11(b) through 1-11(d) show typical ohmmeter circuits for various full-scale resistance ranges.

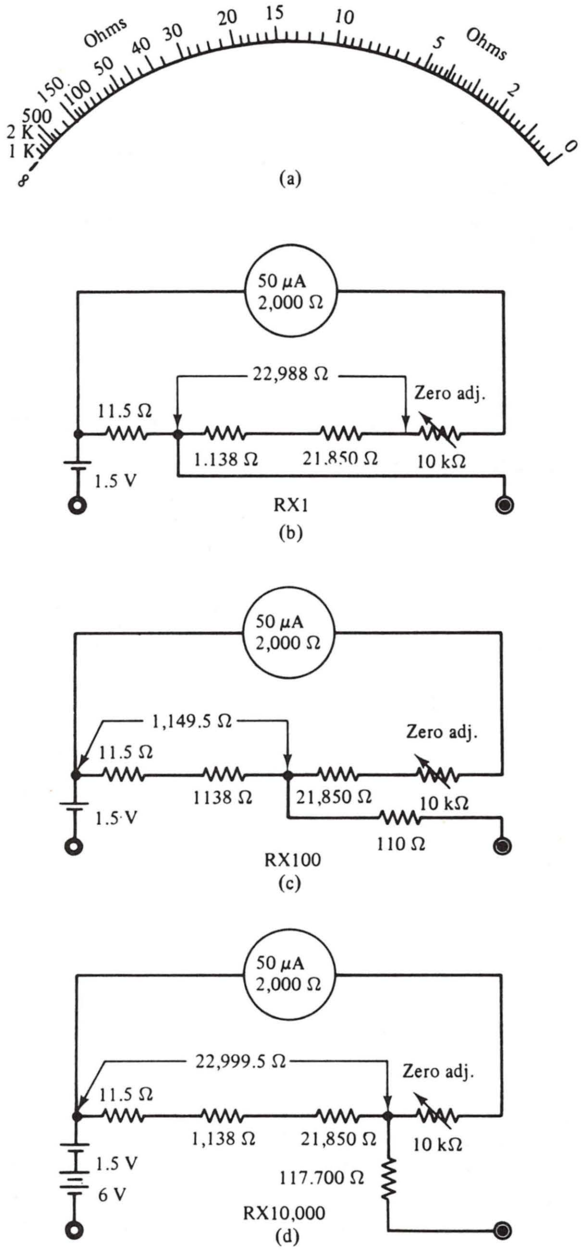


Figure 1-11. (a) Sample ohmmeter scale. (b) Ohmmeter circuit. (c) Ohmmeter circuit. (d) Ohmmeter circuit.

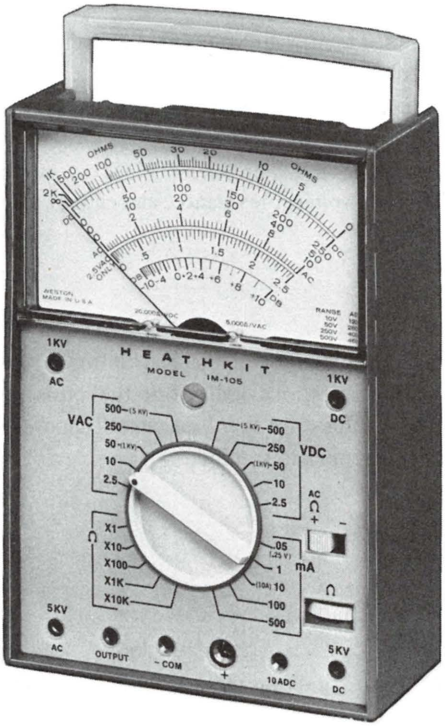


Figure 1-12. Volt-ohm-milliammeter. (Courtesy of the Heath Co.)

1-14 MULTIMETERS

A multimeter (Fig. 1-12) is an instrument that has several switch-selectable voltage, current, and resistance ranges. The main selector switch connects the shunt, multiplier, or range resistors as required. Most instruments also include a rectifier to allow reading ac values (see Chapter 2).

The user must follow the same rules as in individual current and voltage meters, but is warned to be very careful. A typical multimeter might have dc voltage ranges from 0.5 Vdc full scale to 1500 Vdc full scale, so there is always a possibility of destruction of the movement if the rules are not followed.

In general, always store a *volt-ohm-milliammeter* (VOM) in the *off* position, or alternatively, the *highest* voltage range. When using a VOM, begin with the highest range and work down until a readable deflection is obtained.

1-15 SUMMARY

1. The two basic dc meter movements are the *D'Arsonval* and the *taut-band*. Both are examples of PMMC galvanometers.

2. The range of a current meter can be extended by using a *shunt resistor*.
3. A dc current meter can be used to measure voltage by connecting a *multiplier resistor* in series with the meter.
4. An ohmmeter can be made from a dc current meter if a voltage source is provided.
5. A multimeter, called a volt-ohm-milliammeter or VOM, is an instrument that has switch-selectable voltage, resistance, and current ranges.

1-16 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread the appropriate section.

1-17 QUESTIONS

1. An ammeter is connected in ____________ with the load.
2. A voltmeter is connected in ____________ with the load.
3. Name *two* types of PMMC galvanometer meter movements: ____________ and ____________.
4. The same basic electrical principles govern the behavior of PMMC meters and ____________.
5. The range of an ammeter, milliammeter, or microammeter can be increased if a ____________ is used.
6. A voltmeter can be made from a dc current meter movement if a ____________ is connected in ____________ with the meter.
7. Draw a simple ohmmeter circuit.
8. A current meter should have a resistance that is ____________ compared with circuit resistances.
9. A voltmeter should have a resistance that is ____________ compared with the circuit resistances.
10. An ohmmeter may be used in circuits with the power turned on. True or false?

1-18 PROBLEMS

1. You have a 0 to 50 μA dc current meter that has an internal resistance of 4500 ohms. What value of shunt resistor is needed to make the meter read 0 to 500 μA?

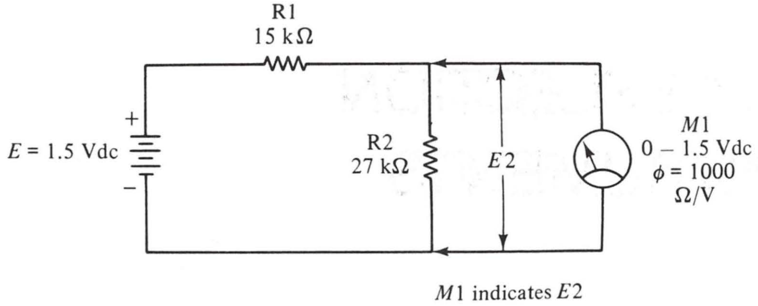


Figure 1-13.

2. A dc current meter is in series with a 12-kΩ load that is connected across a 9-Vdc battery. How much current flows in the meter, if it is assumed that both meter and battery are ideal?
3. A battery has an open-circuit potential of 12 volts dc, and an internal resistance of 22 ohms. A 0 to 1 mA dc meter with an internal resistance of 100 ohms is accidentally (!) connected across this battery. How much current flows?
4. Find the value of a shunt resistor required to make a 0 to 100 μA dc meter read 0 to 1 mA if the meter's internal resistance is 1100 ohms. Use (a) Ohm's law method, and (b) current divider method.
5. Work Problem 4 for a 0 to 20 μA meter. Show your work.
6. What value of multiplier resistor will make a 0 to 100 μA dc meter with an internal resistance of 1300 ohms read 0 to 100 volts?
7. Calculate the multiplier resistance needed to make the meter in Problem 6 read 0 to 2 volts. Show your work.
8. What is the sensitivity of a voltmeter constructed from a 0 to 30 μA movement?
9. Calculate the impedance of a dc voltmeter on the 0-to-50-volt scale if the sensitivity is 20,000 ohms/volt.
10. Refer to Fig. 1-13. (a) What voltage should be indicated on $M1$? (b) What voltage *is* indicated on $M1$? (c) How much percentage of error is there in this measurement?

chapter 2

AC DEFLECTION INSTRUMENTS

2-1 OBJECTIVES

1. To learn the principles of operation of *passive* ac meter movements.
2. To learn the operation of rectifier ac instruments.
3. To learn the *nature* of the value indicated by the different types of instrument.
4. To learn the limitations of each type of meter.

2-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. What is an *electrodynamometer?*
2. Describe an *iron-vane* meter movement.
3. What type of rectifiers are used to make an rf voltmeter?
4. What ac value is read by a rectifier-type ac voltmeter? How is the *scale* of such a meter *usually* calibrated?
5. Describe a conventional ac watt-meter.

2-3 ALTERNATING CURRENT METERS

Neither type of dc PMMC meter movement will correctly indicate the value of an alternating current. There are, however, four basic types of meter that will indicate ac values: *thermocouple, hot-wire, dynamometer,* and *iron-vane*.

2-4 WHICH AC VALUE?

In direct current measurements there is only a single current and voltage measurement, but in alternating current, we might mean any of *several* values when we say, "100 volts ac." We could, for example, mean 100 volts *peak* (E_p), 100 volts *peak-to-peak* ($E_{p\text{-}p}$), 100 volts *rms* (E_{rms}), or 100 volts *average* (E_{av}).

If the ac waveform is a *sine wave,* then we can easily obtain any of the other values if just *one* is known:

$$E_{rms} = 0.707E_p$$
$$E_{p\text{-}p} = 2 \times E_p$$
$$E_{av} = 0.637E_p$$

But these relationships only hold true for a sine wave; if a more complex waveform is being measured, then the relationships will be different because of the amplitude and phase relationships of the *harmonics*. Some ac meters will read *rms* values, while others read *average* values. The user of meters must be aware of which is the case for the meter being used, or else it will prove impossible to correctly *interpret* results.

2-5 THERMOCOUPLE CURRENT METERS

Figure 2-1 shows schematically the *thermocouple ammeter*. This instrument consists of a nonreactive (i.e., resistive) *heating element R* coupled to a *thermocouple* so that both are in the same thermal environment.

A thermocouple is made by joining together two *dissimilar* metals in a *vee-junction* (Fig. 2-1). This assembly is a heat-to-voltage transducer; when the

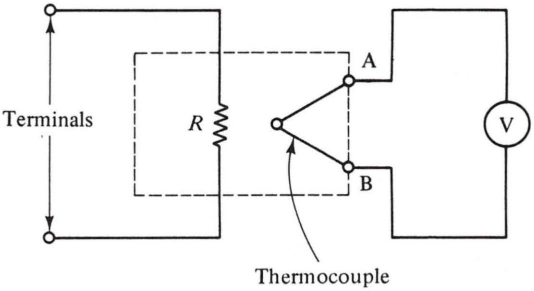


Figure 2-1. Thermocouple meter circuit.

junction is *heated,* a *voltage* appears across terminals A-B. This phenomenon occurs because of a difference in the *work functions* of the two metals forming the thermocouple. Although this voltage is nonlinear at the extremes, it proves reasonably linear in the center portion of its range. Note, however, that zero-referenced thermocouple meters are *not* linear in the low end of their scales. It is necessary to select an instrument that will deflect to the upper half of its range for the current expected in the circuit.

The ac current being measured flows through heater resistance *R* and causes it to become heated, thereby causing the thermocouple to produce an output voltage. A *voltmeter* across terminals A-B measures this millivolt-level potential, but is calibrated in *amperes,* or milliamperes. This operation tells us something of the nature of thermocouple ammeter measurements: They represent the *rms* value of the current.

One method for determining the rms value of any ac waveform is to compare its effect on a heater element with the level of dc required to do the same amount of work. In the classical experiment the ac waveform heats a resistor. When the resistor comes to equilibrium, its temperature is measured. After the resistor has cooled back to ambient, a *calibrated* dc source is connected to the resistor. The dc current is increased until the same temperature is reached. The dc current that caused the same heating as the unknown ac waveform has the same level as the *rms* value of the ac waveform. The thermocouple ammeter, then, since it is based on resistor heating, measures the rms value of the ac waveform.

Almost any ac meter can be used to measure the rms value of a sine wave because of the well-known relationships given in Sec. 2-4, but on nonsinusoidal waveforms these relationships are sometimes difficult to predict. Only an instrument such as the rms-reading thermocouple ammeter can accurately determine the rms value of such waveforms.

2-6 HOT-WIRE AMMETERS

Figure 2-2 shows the *hot-wire* ac meter movement. A *resistance wire* is stretched between two terminals, and is pulled taut by a *tension spring.* The pointer is mounted to a *pivot,* and is tied to the string and spring. A current passing through the resistance element causes it to heat, and thereby expand. When the wire expands, it deflects the meter *pointer;* the amount of deflection is proportional to the rms value of the current.

2-7 THE ELECTRODYNAMOMETER

The *electrodynamometer* is among the oldest ac instruments still in common use. It was invented by the Siemens brothers in Germany following the development of a similar but less effective device by Weber in 1843. By 1910 the configuration

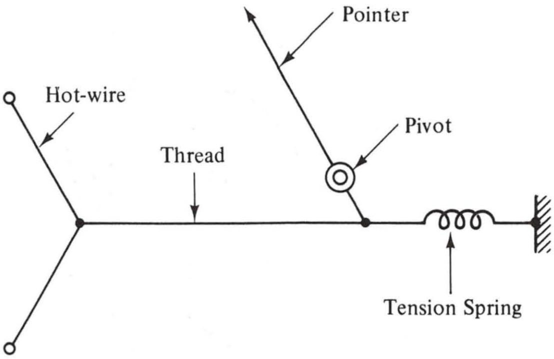


Figure 2-2. Hot-wire meter movement.

of present instruments had been determined, and these designs are still in use today.

The basic electrodynamometer is shown in Fig. 2-3(a). This type of meter differs from the PMMC types in that it creates its own magnetic field from the current flowing in the windings of coils $L1$ through $L3$. Coils $L1$ and $L2$ are *stationary,* while coil $L3$ is free to move, and is attached to the pointer. The *sense* of the windings is such that the magnetic fields of $L1$ and $L2$ will interact with that of $L3$ to create a rotational force on $L3$. The movable coil (i.e., $L3$) rotates an amount that is proportional to the strength of the current flowing in the coils.

The electrodynamometer is a square-law device. That is to say, the pointer deflects to a position that is proportional to the *square* of the *average* current (i.e., i_{av}^2). In most electrodynamometers used to measure current, the dial scale is actually calibrated in terms of the rms current, which is defined as the *square root* of the current i_{av}^2 in sine waves.

It takes a considerable amount of current to deflect the pointer, so most electrodynamometers have a full-scale rating of 100 mA or more, and many are in the 0–5 ampere range.

Electrodynamometer voltmeters can be built from a multiplier resistor and a 100 mA movement, but these can be used only in low impedance applications (e.g., power mains measurements), because the sensitivity is only 10 Ω/V, which is far too low for most electronic circuit applications.

Extending the current range of an electrodynamometer is usually done with a *current transformer* because suitable shunt resistors would be very low-valued, and that renders them difficult to calibrate accurately.

One principal application for electrodynamometers is in the watt-meter used in 60-Hz ac power systems. An example is shown in Fig. 2-3(b). The *stationary* coils are designed to carry a heavy current and are connected in *series* with the load. The movable coil, on the other hand, is wound of a much finer wire, so will carry a much smaller current (100 mA or less). The small coil ($L3$)

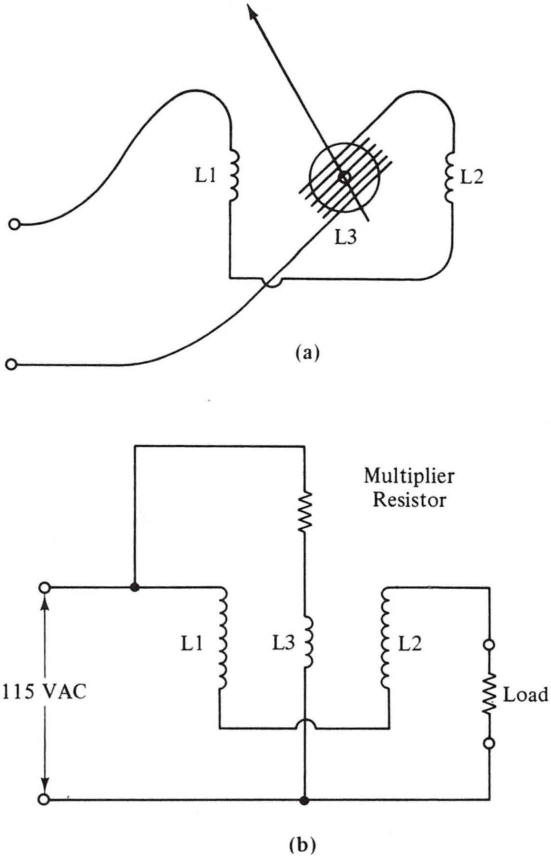


Figure 2-3. (a) Electrodynamometer circuit. (b) Electrodynamometer used as a watt-meter.

is connected *across* the load and serves as a *voltmeter element*. $L3$ may or may not have a multiplier resistor connected, depending upon the design. In the case where there is no multiplier resistor the resistance of the $L3$ winding is sufficient to limit current.

The deflection of the pointer connected to $L3$ is proportional to the product of volts times amperes (V × A), which is given as *watts,* or units of power in nonreactive circuits.

2-8 IRON-VANE MOVEMENTS

The *iron-vane* movement (Fig. 2-4) consists of a coil of wire to carry the current being measured and two iron vanes: one movable and the other stationary. A pointer is attached to the movable vane.

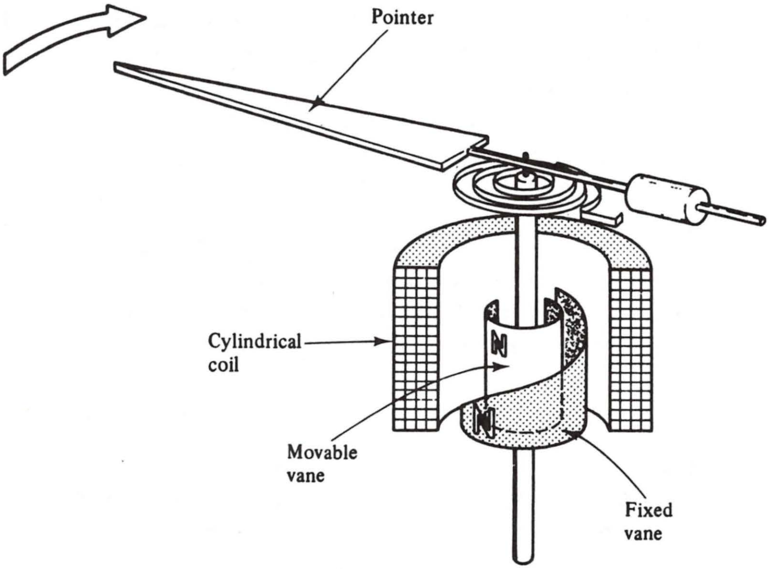


Figure 2-4. Construction of an iron-vane meter movement.

When current flows through the coil it *magnetizes* the two vanes to the *same* polarity, and since *like* poles *repel* each other, the movable vane tends to deflect an amount proportional to the rms value of the current.

At power mains frequencies (50–70 Hz) the iron-vane meter produces accurate readings, but as the frequency of the current is increased *hysteresis* and *eddy current* losses tend to introduce nonlinearity.

At dc there is another type of error: residual magnetism of the vanes by the current in the coil. The dc current flows in only *one* direction, so the vanes will polarize in that direction. Alternating current, on the other hand, changes direction on each half-cycle, so the magnetic field is constantly changing, thereby eliminating buildup of a residual field.

2-9 RECTIFIER INSTRUMENTS

Alternating current *voltage* can be measured on a dc current meter that is connected in series with a multiplier resistor and a rectifier diode. An example is shown in Fig. 2-5(a).

In Fig. 2-5(a) a 10-V rms sine wave is applied across the circuit. The dc meter reads the *average* (not rms) value of the current flowing in the circuit. Note that this current is *not* true dc, but is a pulsating dc waveform. It is *unidirectional,* but does not possess all of the other attributes of dc. As a result, the average current that deflects the meter is equivalent to that which would be generated by a dc voltage across the terminals equal to 0.636 times the peak voltage, or in this example about nine volts.

The half-wave rectifier produces a current flow to the meter over only *half* of the ac cycle, but the meter tends to average the current over the entire cycle. The current indicated on the meter, then, will be *half the average current for one pulse*. As a result, the average potential indicated by the meter used in the circuit of Fig. 2-5 is $\frac{1}{2} \times 0.636 \times E_p$, or in this case $\frac{1}{2}(9\text{ V}) = 4.5$ V.

The average current through the meter can be interpreted as a voltage because of the multiplier resistor and Ohm's law, but this reading is 45 percent of the rms voltage. The meter current is an average of

$$I_m = 0.45(E_{rms}/R1) \qquad (2\text{-}1)$$

where I_m = the meter current in amperes (A)
E_{rms} = the rms value of the potential applied to the input
$R1$ = the value in ohms of the multiplier resistor

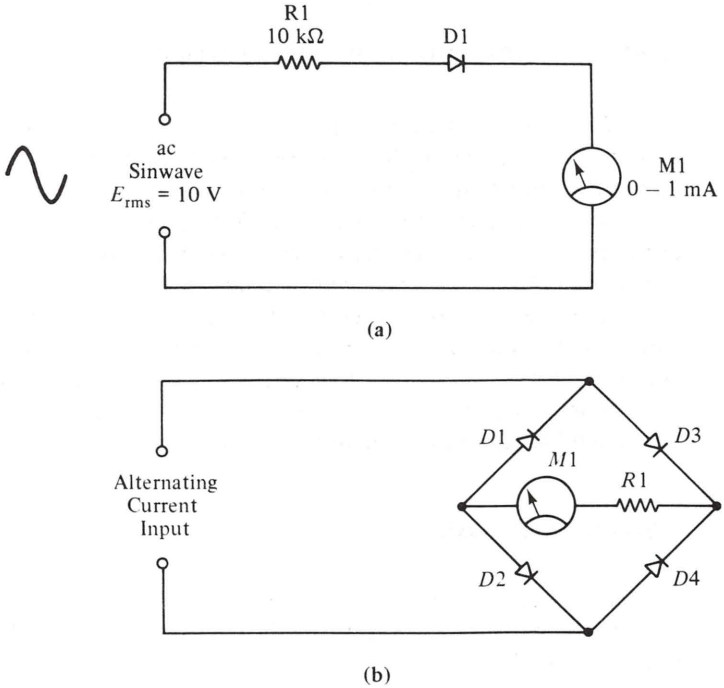


Figure 2-5. (a) Simple ac meter made from a dc movement. (b) Full-wave ac meter made from a dc movement.

Equation (2-1) holds true for half-wave rectified sine waves. Most manufacturers calibrate the dial scale in terms of rms voltage, since that is the most frequently required measurement. In a half-wave rectified meter the *scale* will *read* 1/0.45, or 2.22, times *higher* than the indicated voltage. Note that this calibration is only valid for sine waves, and for other waveshapes a *correction factor* must be used.

Figure 2-5(b) shows the use of a full-wave bridge rectifier with a dc milliammeter. The *average* voltage that is read on such a meter will be $0.9E_p$, so to read rms voltage the scale will have markings that are 1/0.9, or 1.1, times higher than the actual reading. Note again that this calibration applies only to sine waves.

The *ratio* between the *rms* and *average* values is 0.707/0.636, which is the 1.1 scale factor used above.

2-10 TYPES OF RECTIFIERS

The traditional *instrumentation rectifier* used to make dc meters read ac values is a *copper-oxide* diode, sealed in a Bakelite housing. This type of rectifier is suitable for use at ac frequencies well into the audio range, but above the audio range it suffers from nonlinearity problems.

A linearity problem also exists at low applied voltages. This problem is caused by the junction potential of the copper-oxide diode. As a result, most copper-oxide instrumentation rectifiers are used only on ac voltmeters with a full-scale range of 0–1.5 V ac or more.

Alternating current voltmeters that operate into the rf range can be built from dc meter movements and silicon or germanium diodes. Typical types are 1N34 and 1N60 in Ge, and 1N914 and 1N4148 in Si. Instruments that are useful to over 500 mHz can be built with such diodes. The sensitivity of these instruments tends to be better than copper-oxide types because the junction potentials are smaller, e.g., 0.6 volt for Si and 0.2 volt for Ge.

2-11 OTHER AC INSTRUMENTS

Alternating current values can be measured on the oscilloscope, audio voltmeters, and electronic multimeters.

The cathode ray oscilloscope (CRO) is discussed in detail in Chapter 8, so it will not be further discussed here except to state that it allows the user to determine not only the *peak* and peak-to-peak values, but also allows examination of the *waveshape*.

The audio voltmeter (Fig. 2-6) is a vacuum tube or solid-state rectifier instrument preceded by an *amplifier*. This amplifier greatly increases the sensitivity

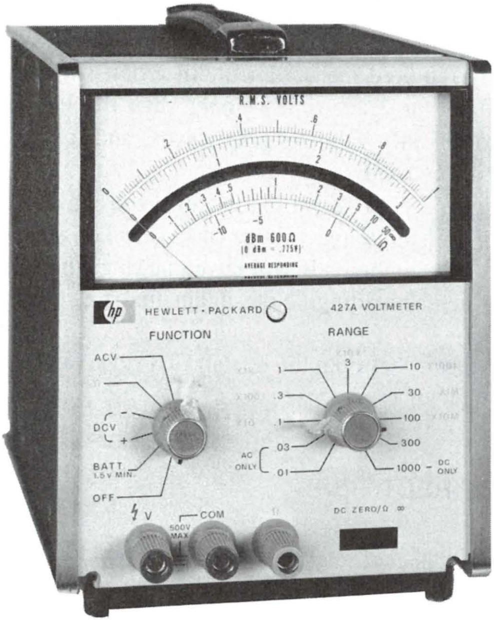


Figure 2-6. AC voltmeter. (Courtesy of Hewlett-Packard.)

of the meter, allowing signals as low as −70 dbm to be measured (*note:* 0 dbm is defined as 1 mW of 1 kHz energy dissipated in a 600-ohm load).

Electronic voltmeters, of which the audio voltmeter is a subclass, may or may not have amplifiers to boost low-level signal, but do use electronic circuits to increase the meter sensitivity (ohms/volt).

The ac voltmeter section of such an instrument may be a rectifier type, or it may use a more complex circuit called an *rms-to-dc converter* that actually computes the rms value of the voltage regardless of the waveform.

2-12 SUMMARY

1. There are four basic types of ac meter movement: *thermocouple, hot-wire, electrodynamometer,* and *iron-vane.*
2. Direct current meters used with *rectifiers* can be used to measure ac values.
3. Different indications are possible from different types of ac instrument used to measure the same potential, because some instruments read *rms,* some *average;* one type reads the *average* of the *current squared.*
4. Correction factors may be used with most ac meters when nonsinusoidal values are measured.

2-13 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread the appropriate section.

2-14 QUESTIONS

1. A thermocouple ammeter indicates the ____________ value of the current waveform.
2. Thermocouple ammeters are used at frequencies up to ____________ mHz.
3. A *hot-wire* ammeter reads the ____________ value of the current waveform.
4. A ____________ is an ac ammeter constructed from a movable coil in the magnetic fields of two stationary coils. All three coils are connected in ____________ with each other.
5. The ammeter in Question 4 is often calibrated in rms values, but is actually reading ____________.
6. A ____________ is used to extend the current range of an electrodynamometer.
7. *Iron-vane* meters read the ____________ value of the ac current waveform.
8. What happens to the iron-vane meter when dc is measured? Can you think of a method for overcoming this problem?
9. A rectifier meter reads the ____________ value of the ac current or voltage waveform.
10. A ____________ diode is often used as the instrumentation rectifier in audio frequency ac meters.

2-15 PROBLEMS

1. A thermocouple ammeter is used to measure a 5-mHz sine wave, and it indicates a current flow of 2.5 amperes in a pure resistance. What is the *peak current* in this waveform?
2. An electrodynamometer is used to measure a sine wave current, and indicates 1.4 amperes rms. What is the *average* value of this current?
3. An iron-vane meter indicates 450 mA when used to measure a 60-Hz ac waveform. What is the rms value of the applied sine wave?
4. A full-wave rectified 60-Hz sine wave has a peak current (I_p) of 1.35

amperes. What will an iron-vane meter read if it is used to measure this current?

5. A half-wave rectified 60-Hz sine wave has a value of 600 mA when measured on an iron-vane meter. What is its *peak* value?
6. A thermocouple RF ammeter used to measure a 60-Hz square-wave current measures 1.2 amperes. What is the average value?

chapter 3

AC AND DC BRIDGES

3-1 OBJECTIVES

1. To learn the principles behind the Wheatstone bridge.
2. To learn the different types of ac bridge.
3. To become familiar with the different types of ac null detector.
4. To learn the applications for the different bridge types.

3-2 SELF-EVALUATION

Before studying the material in this chapter try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Write the null condition equation for a dc Wheatstone bridge.
2. Draw the circuit for a Maxwell bridge.
3. What types of null detectors are suitable for use in an ac bridge?
4. What type of bridge is *best* suited for measurement of (a) high-Q and (b) low-Q components?

3-3 DC WHEATSTONE BRIDGES

Figure 3-1(a) shows the classic dc Wheatstone bridge circuit. The circuit is redrawn in Fig. 3-1(b) to more clearly show the fact that we may consider this circuit as two resistor voltage dividers connected in parallel across a voltage source E. Current $I5$ through the galvanometer is given by

$$I5 = E_0/(R5 + R_m) \tag{3-1}$$

where $I5$ = the current flowing in meter $M1$ in *amperes* (A)
E_0 = the potential in *volts* (V) between points B and D
$R5$ = the resistance of resistor $R5$ in ohms (Ω)
R_m = the resistance of the meter movement in ohms (Ω)

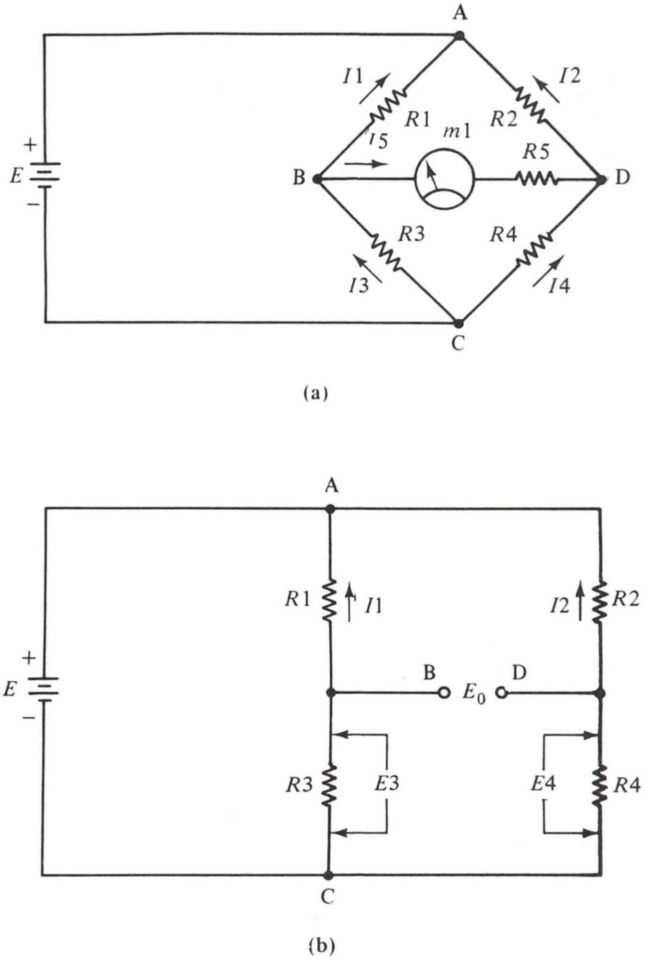


Figure 3-1. (a) The Wheatstone bridge. (b) Redrawn to make more clear.

Voltage E_0 is the difference between voltages $E3$ and $E4$:

$$E_0 = E3 - E4 \tag{3-2}$$

But $E3$ and $E4$ are given in terms of E and the resistances

$$E3 = \frac{ER3}{R1 + R3} \tag{3-3}$$

and $$E4 = \frac{ER4}{R2 + R4} \tag{3-4}$$

By substituting Eqs. (3-3) and (3-4) into Eq. (3-2) we obtain

$$E_0 = E \times \left[\frac{R3}{R1 + R3} - \frac{R4}{R2 + R4} \right] \tag{3-5}$$

Assuming that R_m is negligible, and substituting Eq. (3-5) into Eq. (3-1), we have

$$I5 = \left[\frac{E}{R5} \right] \left[\frac{R3}{R1 + R3} - \frac{R4}{R2 + R4} \right] \tag{3-6}$$

Example 3-1

A dc Wheatstone bridge [Fig. 3-1(a)] uses a 12-volt battery for excitation, and has the following resistance values in the arms: $R1 = 1.2\ \text{k}\Omega$, $R2 = 1.5\ \text{k}\Omega$, $R3 = 4\ \text{k}\Omega$, $R4 = 3.6\ \text{k}\Omega$, and $R5 = 1\ \text{k}\Omega$. Calculate the meter current.

SOLUTION

$$I5 = \frac{E}{R5} \left[\frac{R3}{R1 + R3} - \frac{R4}{R2 + R4} \right] \tag{3-6}$$

$$I5 = \frac{12}{1000} \left[\frac{4}{(1.2 + 4)} - \frac{3.6}{(1.5 + 3.6)} \right]$$

$$I5 = \frac{(12)(0.77 - 0.71)}{1000} = 7.2 \times 10^{-4}\ \text{A} = \mathbf{0.72\ mA}$$

3-4 BRIDGES IN THE NULL CONDITION

The *null condition* in a Wheatstone bridge exists when the output voltage E_0 is zero. We may find the sole necessary condition for null by setting Eq. (3-6) equal to zero.

$$\frac{E}{R5} \times \left[\frac{R3}{R1 + R3} - \frac{R4}{R2 + R4}\right] = 0 \tag{3-7}$$

For Eq. (3-7) to evaluate to zero, either the quantity $E/R5$ or the quantity inside the braces must be zero. The former quantity is never zero in practical situations, so we may conclude that

$$\frac{R3}{R1 + R3} - \frac{R4}{R2 + R4} = 0 \tag{3-8}$$

from which we may also conclude that

$$\frac{R3}{R1 + R3} = \frac{R4}{R2 + R4} \tag{3-9}$$

Equation (3-9) can be simplified by the following steps. At *null,* E_0 is zero, so $I5$ is *also* zero. Therefore,

$$I1 = I3 \tag{3-10}$$

$$I2 = I4 \tag{3-11}$$

We may then write

$$I1R1 = I2R2 \tag{3-12}$$

and

$$I1R3 = I2R4 \tag{3-13}$$

The sole necessary condition for bridge null is obtained by dividing Eq. (3-13) into Eq. (3-12), which yields Eq. (3-14) below.

$$\frac{R1}{R3} = \frac{R2}{R4} \tag{3-14}$$

From Eq. (3-14) we can compute any resistance value if the other three values are known. Note that it is *not* necessary for the resistor values to be equal, only that the *ratios* $R1/R3$ and $R2/R4$ be equal.

Example 3-2

A Wheatstone bridge has the following arm values: $R1 = 1.6\ \mathrm{k}\Omega$, $R2 = 500\Omega$, $R3$ is unknown, and $R4 = 1800\Omega$. What value of $R3$ will bring the bridge into the null condition?

SOLUTION

$$R3 = R1R4/R2 \qquad \text{[(from Eq. (3-14)]}$$

$$R3 = \frac{(1600)(1800)}{500} = \mathbf{5{,}750\ \Omega}$$

3-5 DC BRIDGE APPLICATIONS

Equation (3-14) gives us a means for measuring resistances very accurately—far more accurately, in fact, than is possible on a volt-ohm-milliammeter. In Fig. 3-2, the unknown resistance R_x is given by

$$R_x = \frac{R2R3}{R1} \tag{3-15}$$

(when the bridge is in null).

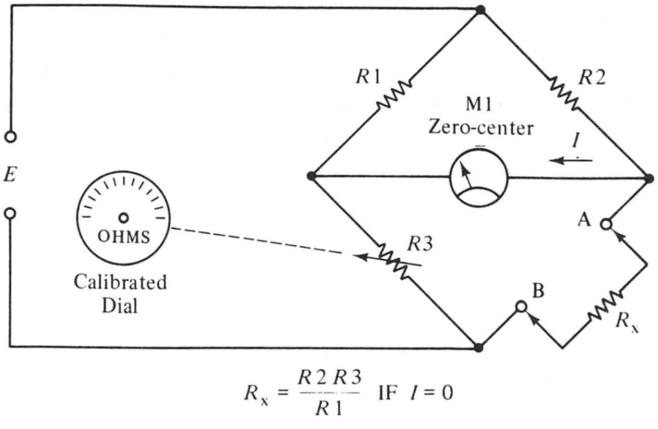


Figure 3-2. Measuring resistance with a Wheatstone bridge.

If we make $R1$ and $R2$ fixed, and link potentiometer $R3$ to a dial that is calibrated in ohms, then we may measure R_x. We could, for example, make $R1 = R2$, so the ratio $R2/R1$ is unity. In that case R_x will equal $R3$ at the null point. To measure R_x, connect it to the bridge terminals A-B and adjust $R3$ for a *zero* reading on meter $M1$. In some commercial bridges resistors $R2/R1$ serve as a range *multiplier,* and have different values on different resistance ranges.

Another category of Wheatstone bridge applications begins with a balanced bridge, and then measures the *amount* of unbalance when one or more of the resistance arms *changes.* This principle is used in many *transducers* in which a stimulus parameter, such as force, temperature, displacement, etc., changes the value of one or more arms in the bridge. If all arms are equal under the condition of zero stimulus, then E_0 will be directly proportional to the value of the stimulus.

3-6 DC NULL INDICATORS

If the bridge is configured such that E_0 is always positive or negative, then any dc meter of appropriate range may be used for $M1$. If, on the other hand, the polarity of the bridge output voltage may swing either positive or negative, then meter $M1$ must be a *zero-center galvanometer.*

Interestingly enough, either a voltmeter or a current meter (milli- or microammeter) may be used for $M1$. In bridges that are passive, a current meter is generally used. Typical ranges fall within the 0–100 microampere to 0–1 milliampere range. The smaller full-scale current movements are capable of better resolution, *provided* that the minimum change of resistance step used·to null the bridge produces current changes that are considerably less than the full-scale current of the meter movement.

Figure 3-3 shows a battery operated dc null meter that is capable of detecting E_0 values down to the *microvolt* range. This instrument is a very sensitive amplified voltmeter that is a special case of the instruments discussed in Chapter 7.

3-7 AC BRIDGES

The dc Wheatstone bridge will work with either ac or dc excitation, although in the former case the null detector must be a device that is sensitive to ac. Certain other types of bridge circuit are designed specifically for ac use, and have inductive or capacitive *reactances* in one or more of the bridge arms.

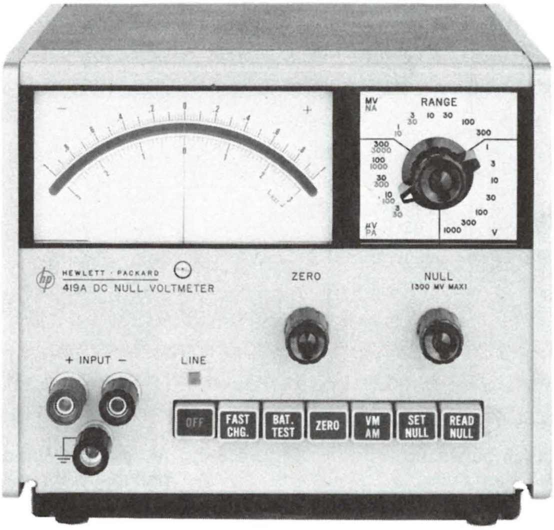


Figure 3-3. Zero-center null voltmeter.(Courtesy of Hewlett-Packard.)

3-8 AC NULL DETECTORS

The dc galvanometer used with the resistive Wheatstone bridge will not respond to ac, so it cannot be used directly as the null detector in ac bridges. Many rectifier ac meters based on the dc movement are not sensitive enough to indicate the null precisely, so are limited to applications where a relatively shallow—i.e., broad—null can be tolerated.

Alternating current electronic voltmeters and oscilloscopes, on the other hand, are quite sensitive, so can indicate the ac null with great precision. These instruments can be used to replace the dc galvanometer-rectifier combination where sensitive measurements are to be made.

In some cases, high-impedance earphones are used as a null detector (see Fig. 3-4) in ac bridges. A surprising degree of sensitivity is possible if the operator has high hearing acuity. In general, highly efficient communications (*not* high fidelity) earphones with an impedance of 1000 ohms or higher are used.

Problems sometimes occur when one is trying to use instruments with ground referenced inputs as a null detector, and these problems are especially troublesome if the ac excitation source is also grounded.

There are two basic approaches to solving this problem: One is transformer coupling to the null detector (thereby isolating the detector from the bridge), and the other is the use of a differential amplifier.

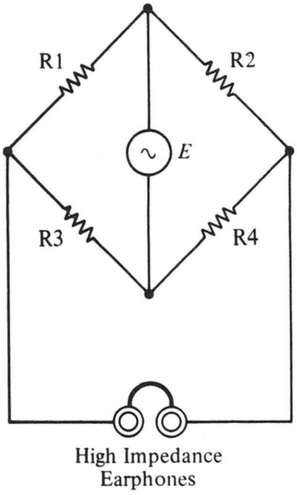


Figure 3-4. AC excited Wheatstone bridge using headphones as the null detector.

Figure 3-5(a) shows the use of a transformer to couple the signal from the E_0 nodes of the bridge to an external null indicator such as an oscilloscope or ac voltmeter. In this particular example the null indicator is an ac voltmeter. Transformer coupling is often used in RF bridges to provide isolation. The bridge components are sensitive to body capacitance from the operator or other nearby disturbances. The bridge then may be located at the site of the measurement in relative isolation from the operator, and the output signal may be brought to the null indicator through a coaxial cable transmission line.

A differential amplifier will produce a single-ended output from a differential input signal. The differential amplifier is useful for isolation, as described above, or for increasing the sensitivity of the null indicator by amplification of the bridge output signal.

Many oscilloscopes can be equipped with a differential amplifier plug-in (see Chapter 8), while even most nonplug-in models can be used in a quasidifferential mode if they are equipped with two channels. To use a two-channel oscilloscope in quasidifferential, use the two inputs as inputs to a differential amplifier, and set the oscilloscope's *mode* switch in *A-B*. Alternatively, an operational amplifier circuit can be constructed by using information presented in Chapter 12.

3-9 PHASE DETECTORS

The ac bridges of the following sections require a sensitive phase detector to indicate the output condition. This detector need not be able to indicate absolute

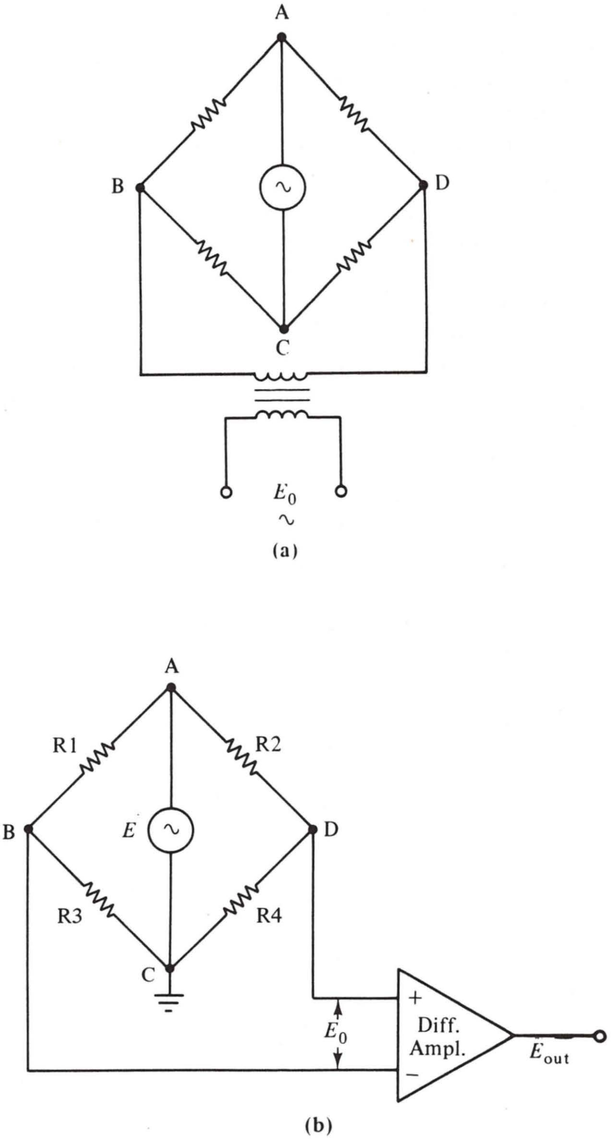


Figure 3-5. (a) Transformer-coupled ac Wheatstone bridge. (b) Differential amplifier output for an ac Wheatstone bridge.

phase—that is, degrees of phase angle θ, but only *relative phase*—that is, whether or not a signal is *in phase* or *out of phase* with a reference signal.

Figure 3-6 shows a typical phase detector circuit. Signal E_{ref} is a sine wave, and is usually the same signal that is used for bridge excitation. This arrangement allows us to measure changes in phase angle θ due to the reactive component in each bridge arm.

Under *no-signal* conditions diode $D1$ will conduct on one-half of the reference signal cycle, while $D2$ will conduct on the alternate half-cycle. The effect of $C1$, $C2$ and the natural mechanical damping of meter movement $M1$ is to *average* these alternations over the entire cycle, producing a net potential across $M1$ of zero.

When an input signal is applied to the circuit, it algebraically adds to E_{ref}, so it *aids* E_{ref} when *in phase* and *opposes* E_{ref} when *out of phase*.

Signal voltage E_o is applied to diodes $D1$ and $D2$ in parallel, while E_{ref} is applied in push-pull. On one-half of its cycle E_o will forward-bias *both* diodes simultaneously, whereas on the opposite half-cycle it will reverse-bias both diodes.

If E_{ref} and E_o are in phase, then the positive half-cycles *coincide*, so with the polarities shown, the anode of $D1$ sees a large positive potential created by the summation of $+E_{ref}$ and E_o, while $D2$ sees a lower potential; $-E_{ref}$ and E_o partially or totally *cancel* [see Fig. 3-6(b)] at $D2$. This situation causes the output of $D1$ to be *more positive* than the output of $D2$, so meter $M1$ swings positive. On the second half-cycle the polarities are reversed, but the output relationships seen by $M1$ are the same.

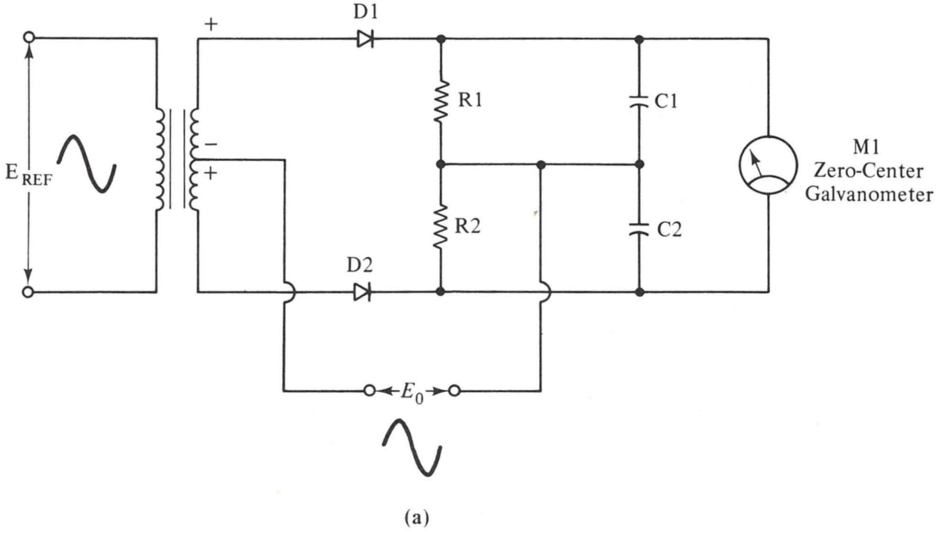


Figure 3-6. (a) Phase detector circuit.

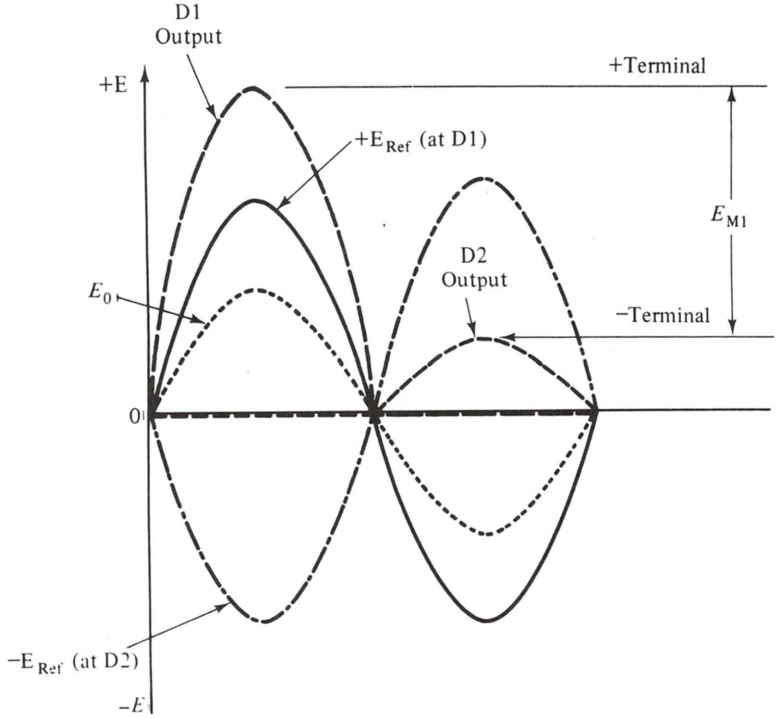


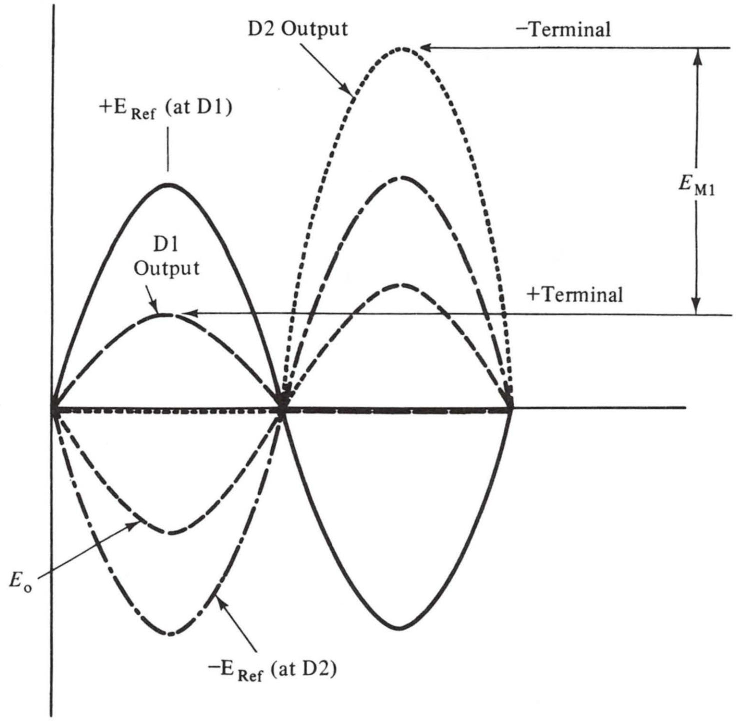
Figure 3-6. *(cont.)* (b) Signal relationships in phase detector(in-phase).

For an out-of-phase input signal E_0, exactly the opposite occurs [Fig. 3-6(c)], so the output of $D2$ is more positive than $D1$, so the meter $M1$ swings negative.

This phase detector, then, responds to the *relative phase* of the input signal by creating a dc output that is positive for an in-phase condition, and negative for an out-of-phase condition. The amplitude of E_{ref} is constant, but the amplitude of E_0 changes with bridge balance conditions, so we may conclude that the circuit of Fig. 3-6(a) is sensitive to *both* phase and magnitude.

3-10 TYPES OF AC BRIDGE

Although the resistive Wheatstone bridge is technically an "ac bridge" if the excitation source is ac, we usually limit the designation to those bridge types that use a reactive element such as a capacitor or inductor in one or more bridge arms. Such bridges may be used for *impedance* measurements. In this chapter we will consider the *Maxwell, Hay,* and *Schering* bridge circuits.



(c)

Figure 3-6. *(cont.)* (c) Signal relationships in phase detector (out-of-phase).

The general form of an ac bridge is shown in Fig. 3-7; it is the standard Wheatstone bridge configuration with the resistance elements in the arms replaced by complex impedances of the general type

$$Z = \sqrt{R^2 + (X_L - X_c)^2} \qquad (3\text{-}16)$$

where Z = the complex impedance in ohms (Ω)
R = the resistance in ohms (Ω)
X_L = the inductive reactance in ohms (Ω)
X_c = the capacitive reactance in ohms (Ω)

The output signal from this type of bridge will have only a *magnitude* component if the inductive and capacitive reactances cancel, but will have both *magnitude* and *phase* components if X_L does not exactly balance X_c.

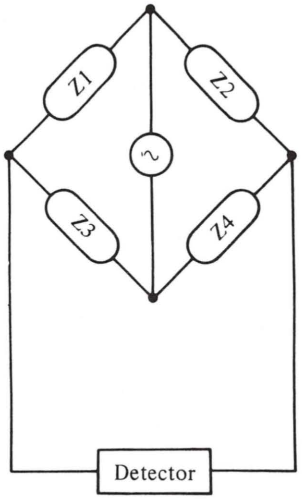


Figure 3-7. Generalized circuit for complex ac bridges.

3-11 MAXWELL'S BRIDGE

Figure 3-8(a) shows the *Maxwell* bridge circuit in which Z2 and Z3 of the general configuration have been replaced by pure resistances, Z1 has been replaced by a parallel combination $R1/C1$, and Z4 by the series combination $L1/R4$.

The null condition for a Maxwell bridge circuit is given by the following equations:

$$L1 = R2 \times R3 \times C1 \tag{3-17}$$

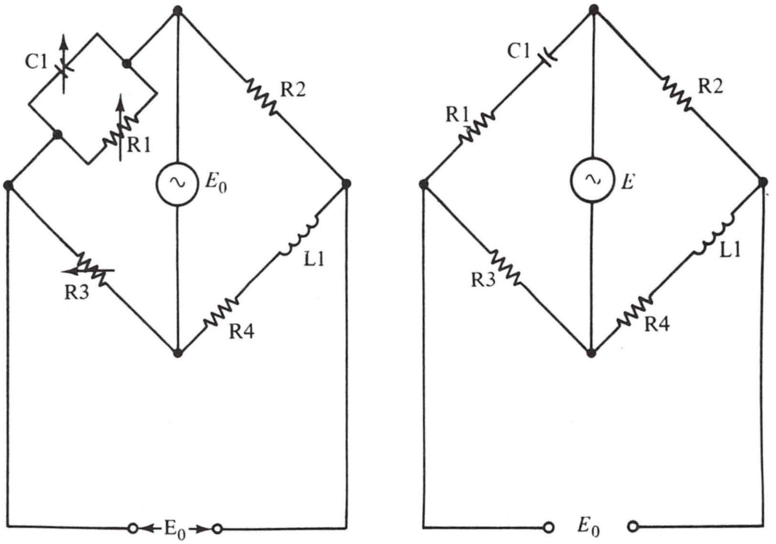
$$R4 = \frac{R2 \times R3}{R1} \tag{3-18}$$

The Maxwell bridge is often used to measure unknown values of inductance (e.g., $L1$) because the balance equations are totally *independent* of frequency. The bridge is also not too sensitive to resistive losses in the inductor. Additionally, it is much easier to obtain easily calibrated capacitor standards for $C1$ than it is to obtain inductor standards for $L1$. As a result, the principal use of the Maxwell bridge is the measurement of inductances.

The Maxwell circuit is often used in *Q-meters,* i.e., instruments that measure the quality factor Q of inductors. The equation for Q, however, is frequency-sensitive:

$$Q_L = \omega \times R1 \times C1 \tag{3-19}$$

where ω is the quantity $2\pi f$ and $R1$ and $C1$ are as previously.



$$L1 = R2\,R3\,C1$$

$$R4 = \frac{R3}{R1} \times R2$$

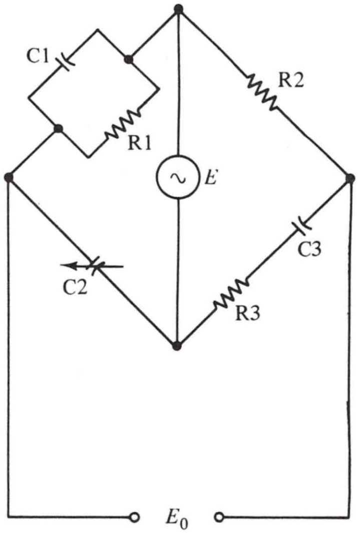
$$Q_L = \omega\,C1\,R1$$

(a)

$$R4 = \frac{R2\,R3}{R1} \times \frac{1}{Q^2 + 1} \qquad L1 = \frac{R2\,R3\,C1}{1 + \left(\frac{1}{Q}\right)^2}$$

$$Q = \frac{1}{\omega\,R1\,C1}$$

(b)



$$R3 = \frac{C1}{C2} \times R2 \qquad C3 = \frac{R1}{R2} \times C2$$

(c)

Figure 3-8. (a) Maxwell's bridge. (b) The Hay bridge. (c) The Schering bridge.

3-12 THE HAY BRIDGE

The Hay bridge circuit [see Fig. 3-8(b)] is physically similar to Maxwell's bridge, except that the $R1/C1$ combination is connected in *series*. The Hay bridge is, however, frequency-sensitive. The balance equations are given below.

$$L1 = \frac{R2 \times R3 \times C1}{1 + \left[\dfrac{1}{Q}\right]^2} \tag{3-20}$$

$$R4 = \left[\frac{R2 \times R3}{R1}\right] \times \left[\frac{1}{Q^2 + 1}\right] \tag{3-21}$$

$$Q = \frac{1}{\omega \times R1 \times C1} \tag{3-22}$$

The Hay bridge is preferred for measuring *inductances* with high Q figures, while Maxwell's bridge works best with low-Q inductors.

Note that a frequency-independent version of Eq. (3-20) is possible when the Q is very high—that is, greater than about ten—in which case Eq. (3-20) can be rewritten in the approximation form

$$L1 = R2 \times R3 \times C1 \tag{3-23}$$

The error caused by using the approximation of Eq. (3-23) is 1 percent if the Q is 10, and 0.01 percent if the Q is 100.

Proof: Let $R2C3C1 = k$ = constant

Assume $Q = 10$

a. [from Eq. (3-20)]

$$L1 = k/(1 + (1/Q)^2)$$
$$L1 = k/[1 + (1/10)^2]$$
$$L1 = k/(1 + 0.01) = k/1.01 = 0.99k$$

b. [from Eq. (3-22)] $L1 = k$

c. $\dfrac{0.99k - k}{0.99k} \times 100 = -1\%$

3-13 THE SCHERING BRIDGE

The Schering bridge circuit is shown in Fig. 3-8(c). This circuit uses a parallel RC network $R1/C1$ for $Z1$, a resistance $R2$ for $Z2$, a capacitive reactance $C2$ for $Z3$, and a series RC network $R3/C3$ for $Z4$. The Schering bridge balance equations are

$$R3 = \frac{R2 \times C1}{C2} \tag{3-24}$$

$$C3 = \frac{C2 \times R1}{R2} \tag{3-25}$$

The Schering bridge is used primarily for the measurement of *capacitance* and the *power factor* of capacitors. In the latter application no actual $R3$ is connected into the circuit, making the series resistance of the capacitor being tested (e.g., $C3$) the only resistance in that arm of the bridge. The capacitor Q is found from

$$Q_{C3} = \frac{1}{\omega \times R1 \times C1} \tag{3-26}$$

3-14 SUMMARY

1. The *Wheatstone* bridge consists of only resistances in the respective branches, and may be used with either ac or dc excitation.
2. The *Maxwell* bridge and *Hay* bridge are used for inductance measurements, whereas the *Schering* bridge is used for capacitance measurement. The Wheatstone bridge is used mainly for resistance measurements.
3. The *zero-center galvanometer* is used as a null detector in dc Wheatstone bridges.
4. Alternating-current current or voltmeters are used as null detectors in ac bridges when there are no phase components in the output signal, while an oscilloscope or phase detector can be used when there are both phase and magnitude components.

3-15 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems below.

Place a mark beside each problem or question that you cannot answer, and then go back and reread appropriate sections of the text.

3-16 QUESTIONS

1. Draw the circuit for a dc Wheatstone bridge and label the components ($R1, R2, R3$, etc.).
2. Using the component designations given in Question 1, write the *null condition* equations.
3. Draw the circuit for a *Maxwell* bridge, label the components, and write the null condition equations.
4. Draw the circuit for a *Hay* bridge, label the components, and write the null condition equations.
5. Draw the circuit for a *Schering* bridge, label the components, and write the null condition equations.
6. What type of ac null detector is *best* suited for (a) the Wheatstone bridge? (b) Hay bridge? (c) Schering bridge? (d) Maxwell bridge?
7. What type of bridge is best suited for measuring capacitance?
8. What type of bridge is best suited for measuring the inductance of high-Q coils?
9. What type of bridge is best suited for measuring the inductance of low-Q coils?
10. What type of bridge is best suited for measuring the power factor of a capacitor?
11. A zero-center galvanometer with a range of 1–0–1 mA is used as the null detector in a dc Wheatstone bridge: What type of circuit or device may be used to increase the *sensitivity* of the galvanometer so that 10 microamperes will deflect the meter full-scale?
12. What type of device or circuit may be used with an ac Wheatstone bridge to drive a ground-referenced null indicator?

3-17 PROBLEMS

1. Is the Wheatstone bridge of Fig. 3-9 balanced? What is the value of E_0?
2. $R4$ in Fig. 3-9 is changed to 5.2kΩ. What is the value of E_0?
3. A 20-kΩ resistor is connected between points A and B in Fig. 3-9, and $R4$ is changed to 4.7 kΩ. Calculate the current through the 20-kΩ resistor.
4. What value would E_0 in Fig. 3-9 become if $R2$ were shunted by a 100-kΩ resistor?

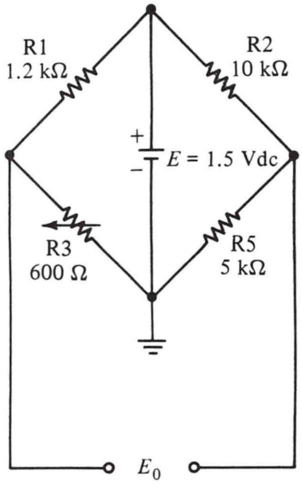


Figure 3-9.

5. An unknown resistance R_x is connected into Fig. 3-9 in place of $R4$, and $R3$ is adjusted to reestablish the null condition. What is the value of R_x if $R3$ has a value of 576 ohms?
6. What is the value of $L1$ in Fig. 3-10?
7. What is the value of $R4$ in Fig. 3-10?
8. Find the Q of the $R4/L1$ network in Fig. 3-10.

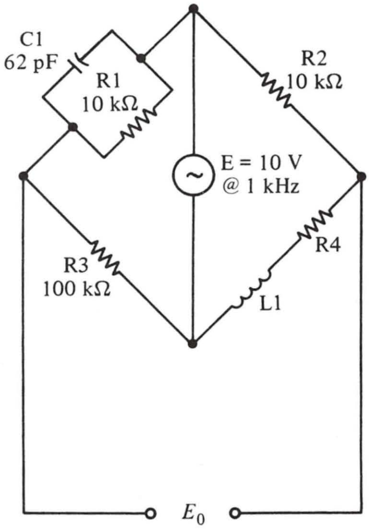


Figure 3-10.

9. A *Hay* bridge is formed by using the *same* components as in Fig. 3-10. An unknown inductor is connected in place of $L1$, and the bridge is adjusted to null. At that point it is found that $C1 = 93$ pF. What is the value of the unknown inductor?
10. What is the Q of the coil in Problem 9?
11. Use the *loop* method of network analysis to derive the expression for E_0 in an *unbalanced* dc Wheatstone bridge.

3-18 REFERENCES

1. F. E. Terman and J. M. Pettit, *Electronic Measurements,* 2nd edition. New York: McGraw-Hill, 1952.
2. Walter Kidwell, *Electronic Instruments and Measurements*. New York: McGraw-Hill, 1969.
3. Sol Prensky, *Electronic Instrumentation,* 2nd edition. Englewood Cliffs, N.J.: Prentice-Hall, Inc., 1971.
4. Stanley Wolf, *Guide To Electronic Measurements and Laboratory Practice*. Englewood Cliffs, N.J.: Prentice-Hall, Inc., 1973.

chapter 4

COMPARISON MEASUREMENTS

4-1 OBJECTIVES

1. To learn the construction and operation of a *precision potentiometer*.
2. To learn how to make *precision measurements* with a potentiometer.
3. To learn the operation of an automatic potentiometer.

4-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Describe a *precision potentiometer* in your own words.
2. How do you measure an unknown *current* using the *equal deflection* technique?
3. What is a Weston cell?

4-3 COMPARISON MEASUREMENTS

A meter movement calibrated to read a voltage or current is not uniformly accurate over its entire range. Most deflection meters produce the best results in the upper half or upper third of their range. High-quality deflection movements are accurate to 1 or 2 percent of full scale in the upper half of the range, whereas lower-quality instruments are accurate to only 5 percent.

Comparison measurements, on the other hand, generally use either a null condition or equal deflection of a meter movement to indicate the point where an unknown voltage or current is *equal* to a precise *reference* source. The Wheatstone bridge (Chapter 3) and *potentiometer* are examples of comparison measurements. In this chapter we will discuss the precision potentiometer. Note that in this application the word "potentiometer" is not used to refer to the low-cost electronic component often used as a radio volume control, but to a precision cousin.

Two techniques are commonly employed in potentiometer measurements: *equal deflection* and *null*. In the equal deflection method the unknown and calibration sources are alternately connected to the meter, and the potentiometer is adjusted so that both produce the *same* deflection on a meter movement. This procedure is more accurate than ordinary deflection methods. Although the accuracy of the meter may be in doubt, it is quite resettable; that is to say, the same potential applied to the meter on different occasions will always produce the same *deflection;* even if the *dial calibration* is in error, the deflection will be the same.

4-4 POTENTIOMETERS

Most electronics students are familiar with the type of potentiometer used to make circuit adjustments, e.g., the radio volume control mentioned above. But these electronic components are only a crude relative of the type of potentiometer used to make comparison measurements. Most of these instrumentation potentiometers consist of a resistance wire or *element* stretched between two terminals [Fig. 4-1 (a)] and a *wiper* that slides along the resistance element, and is connected to a third terminal. If a potential, that is, voltage, is applied across the resistance element, then the potential measured between the wiper and either end terminal is a function of the wiper *position*. If the position can be precisely determined, then the voltage between the wiper and end terminals is also precisely known, hence the name *potentiometer*.

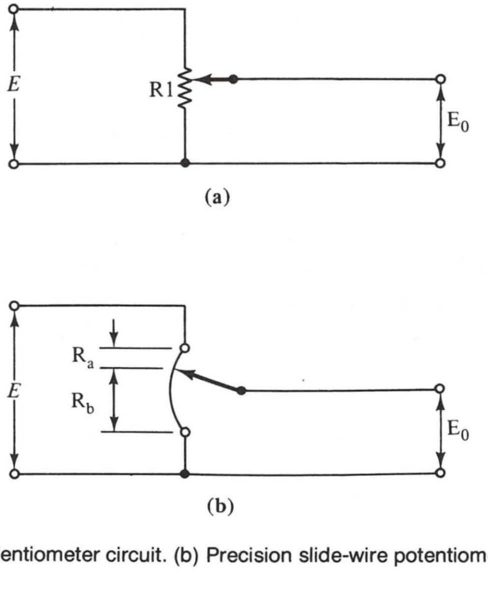


Figure 4-1. (a) Potentiometer circuit. (b) Precision slide-wire potentiometer.

Precision potentiometers usually employ a very precise resistance wire [Fig. 4-1(b)] as the resistance element. If the wire is formed to have a uniform cross-sectional area and composition, then a very accurate (0.01 to 0.05 percent) potentiometer can be made. Even if temperature changes causes resistance changes in the wire, the change is essentially uniform, so it does not affect the output calibration. The output voltage in a nonloaded potentiometer is given as the ratio

$$E_0 = \frac{ER_b}{R_a + R_b} \tag{4-1}$$

where E_0 = the output potential in *volts* (V)
E = the reference potential in *volts* (V)
R_a = the resistance in ohms (Ω) between terminals *a* and *c* [Fig. 4-1(b)]
R_b = the resistance in ohms (Ω) between terminals *b* and *c*

Example 4-1

The overall resistance of the slide wire in Fig. 4-1(b) is 150 ohms. If the wiper is set to the ⅔ mark, then $R_a = 50\ \Omega$ and $R_b = 100\ \Omega$. Find E_0 if $E = 1.36$ volts dc.

SOLUTION

$$E_0 = ER_b/(R_a + R_b) \tag{4-1}$$

$$E_0 = \frac{(1.36\text{ V})(100\ \Omega)}{(50 + 100)\ \Omega}$$

$$E_0 = (1.36\text{ V})(100)/(150) = \mathbf{0.906\ V}$$

If reference potential E is precisely known, and the relative position of the wiper can be determined, then very precise output voltage (E_0) can be created.

Consider what happens when resistance $R_{a\text{-}b}$ changes from 150 ohms to, say, 156 ohms due to a change in temperature. By using the changed resistance in Example 4-1 we produce the *same* output voltage:

$$(R_{a\text{-}b} \rightarrow 156\Omega, \quad \text{so} \quad R_a \rightarrow 52\ \Omega \text{ and } \quad R_b \rightarrow 104\ \Omega)$$

$$E_0 = \frac{(1.36\text{ V})(104\ \Omega)}{(52 + 104)\ \Omega}$$

$$E_0 = (1.36\text{ V})(104)/(156) = \mathbf{0.906\ V}$$

Clearly, the precision potentiometer maintains its accuracy even under changes in temperature, something that *cannot* be said of certain other instruments.

4-5 POTENTIOMETER CIRCUIT

Figure 4-2 shows the circuit for a manual potentiometer used to measure an unknown potential E_x.

Resistor $R1$ is the potentiometer, and consists of a slide-wire element [as in Fig. 4-1(b)], and a *variable* working voltage E. The wiper of $R1$ is mechanically coupled to a dial that measures its position precisely and gives this relative position on a micrometer scale of 0000 to 9999. The indication is 0000 when R_b is zero, and 9999 when R_a is zero (i.e., R_b is full scale). The output potential E_0, then, can be expressed as a decimal fraction of E.

Switch $S1$ is a normally-open pushbutton switch, and is used to turn on the meter circuit. In normal operation $S1$ is just *tapped* by the operator until the meter indicates that the circuit is close to null.

Switch $S2$ is the *sensitivity* control, and is normally open until the meter is near the null. At that time, the operator closes $S2$, increasing the sensitivity, and then completes the adjustment.

Switch $S3$ selects either the unknown potential E_x or the precision reference source E_r to balance against E_0.

There are three voltage sources in this circuit: E, E_x, and E_r. Voltage E is the working voltage. This supply is a variable voltage source. It need not be ac-

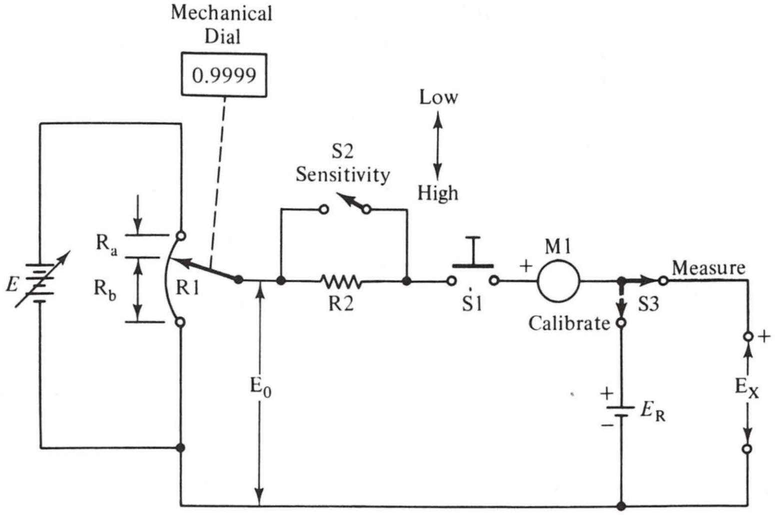


Figure 4-2. Instrumentation potentiometer.

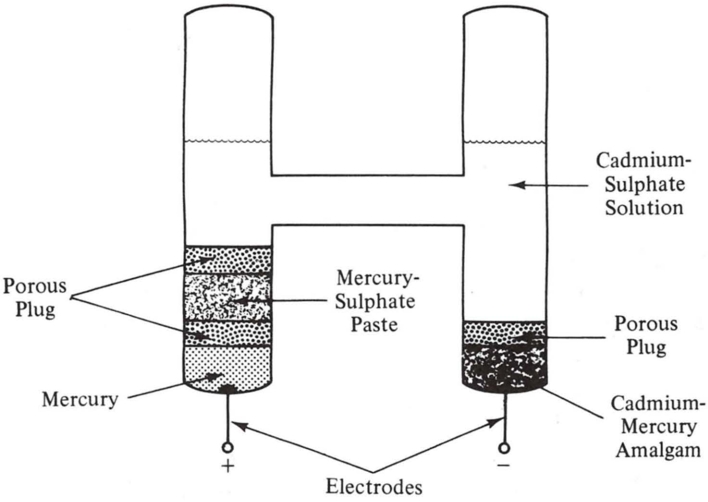


Figure 4-3. Weston cell.

curate, but must be *stable*. In some models the working voltage is a battery connected in series with a rheostat (two-terminal variable resistor), but in many recent models E is a variable electronically regulated power supply.

Source E_r is a precision reference source. In most older, and many recent, models E_r would be taken from a *Weston cell,* also called a *standard cell* (Fig. 4-3).

The Weston cell is available in two styles: *saturated* and *unsaturated.* These terms (see Fig. 4-3) refer to the fact that the solution of cadmium sulphate used in the cell is saturated in one type, but not in the other. The saturated cell is accurate to within 1 *microvolt* of its test value at a temperature of 20°C, but the unsaturated type is considered more portable.

The theoretical output voltage for a Weston cell at 20°C is 1.01830 volts dc, and the potential *changes* at a rate of approximately 40 μV/°C.

The unsaturated cell is built in a closed container, so it can be made portable. It exhibits nearly the same temperature properties as the saturated version, but is known to have a *negative drift* in the terminal voltage of approximately 3 μV/month. Also, at 20°C the output voltage when the cell is new may be any value between 1.018 and 1.020 volts. The actual value for any specific cell is given on a label attached to the cell.

The Weston cell, although very accurate, is a delicate device, and there are several restrictions on its use. The most important restrictions are

1. Never allow more than 10^{-4} amperes to pass through the cell in either charge or discharge modes. This limit restricts the minimum load resistance across the cell to 10.2 kΩ.
2. Always maintain the cell temperature between 4°C and 40°C.

4-6 POTENTIOMETER OPERATION

The instructions for operating the manual potentiometer are given below. Refer to Fig. 4-2 while reading.

Calibration

1. Set $S3$ to *calibrate,* $S2$ to *low* (e.g., $R3$ in-circuit), and potentiometer $R1$ to half-scale (5000) or some other point specified by the manufacturer.
2. *Briefly* tap $S1$ and note whether the pointer on meter $M1$ deflects positive or negative. If
   a. $E_0 > E_r$, then $M1$ swings positive.
   b. $E_0 < E_r$, then $M1$ swings negative.
3. Adjust voltage source E in *small steps* according to Step 2a or 2b until $E_0 = E_r$. If $M1$ deflected positive, then E must be decreased, but if $M1$ deflected in the negative direction E must be increased.
4. Repeat Step 3, using very small incremental changes in E until $M1$ reads nearly zero; then close $S2$—i.e., set $S2$ to *high*—and continue repeating Step 3 until no further improvement is possible.
5. The output of the potentiometer E_0 is now equal to E_r. Set $S3$ to *measure* and $S2$ to *low*.

Use

1. Perform the calibration procedure above.
2. Using the same tapping method, with the unknown voltage E_x connected, adjust potentiometer $R1$ in small steps until $M1$ reads nearly zero.
3. Close $S2$, and repeat Step 2 until no further improvement is possible. Voltage E_x is calculated from

$$E_x = \frac{E_r(R1 \text{ dial reading})}{9999} \tag{4-2}$$

Example 4-2

Find the unknown potential E_x if the reference voltage is 1.0186 V and the micrometer dial reading is 6991.

SOLUTION

$$E_x = E_r(6991)/9999 \tag{4-2}$$
$$E_x = (1.0186 \text{ V})(6991)/(9999)$$
$$E_x = 7.122 \times 10^{-1} \text{ V} = \mathbf{0.7122V}$$

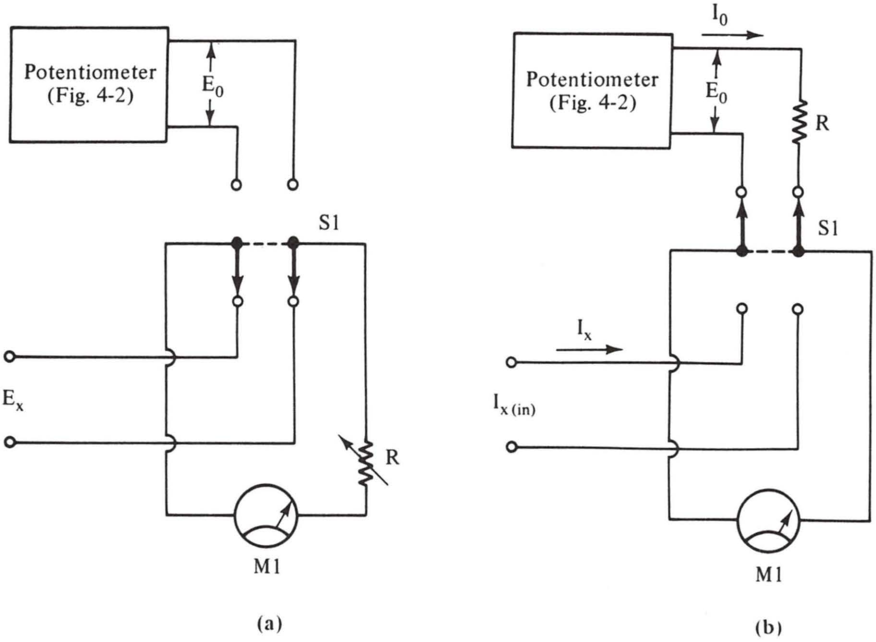


Figure 4-4. (a) Equal deflection measurement of voltage. (b) Equal deflection measurement of current.

4-7 EQUAL DEFLECTION METHODS

An equal deflection method for measuring voltage is shown in Fig. 4-4(a). The potentiometer is the circuit of Fig. 4-2, and the calibration procedure must be followed before a measurement can be made.

Voltmeter $M1$ must have a very high impedance, and must have a full-scale range of approximately, but greater than, E_0, and preferably, several ranges with E_0 being the highest.

Switch $S1$ connects the voltmeter to the unknown voltage *first*. The *range* of the voltmeter is adjusted so that the meter pointer deflects to some specific mark on the scale. Adjust R so that the pointer lands on an easily repeatable point on the upper half of the meter scale.

Next, the switch (e.g., $S1$) is set to connect the potentiometer to the voltmeter. Adjust the output E_0 of the potentiometer ($R1$ in Fig. 4-2) until the pointer of the meter deflects to the *same* mark as E_x. Calculate the unknown voltage from the dial setting of the potentiometer.

A very similar technique is used with the circuit in Fig. 4-4(b) to measure currents. Adjust E_0 so that I_x and I_0 produce the same deflection on meter $M1$. In that case,

$$I_x = I_o$$

and

$$I_0 = E_0/R$$

[R is a precision resistor, and E_0 is calculated from Eq. (4-2).]

4-8 AUTOMATED COMPARISON CIRCUITS

There are two types of automated comparison potentiometers: *servomechanisms* and *electronic*. The servomechanism type is electromechanical, and usually has a pen to record the potential's waveform on graph paper. These machines will be discussed in detail in Chapter 10, so only the electronic type will be discussed in this section.

The null condition is indicated by a *comparator* circuit [Fig. 4-5(a)]. In its most basic form a comparator is a very high gain operational amplifier (see Chapter 12) with no negative feedback. The open-loop gain of most operational amplifiers is at least 100,000, so theoretically a differential potential ($E_r - E_x$) of 10 microvolts will saturate the amplifier output. In actual practice, however, the potential required to saturate the comparator will be between 10 μV and 10 mV, depending upon the model selected.

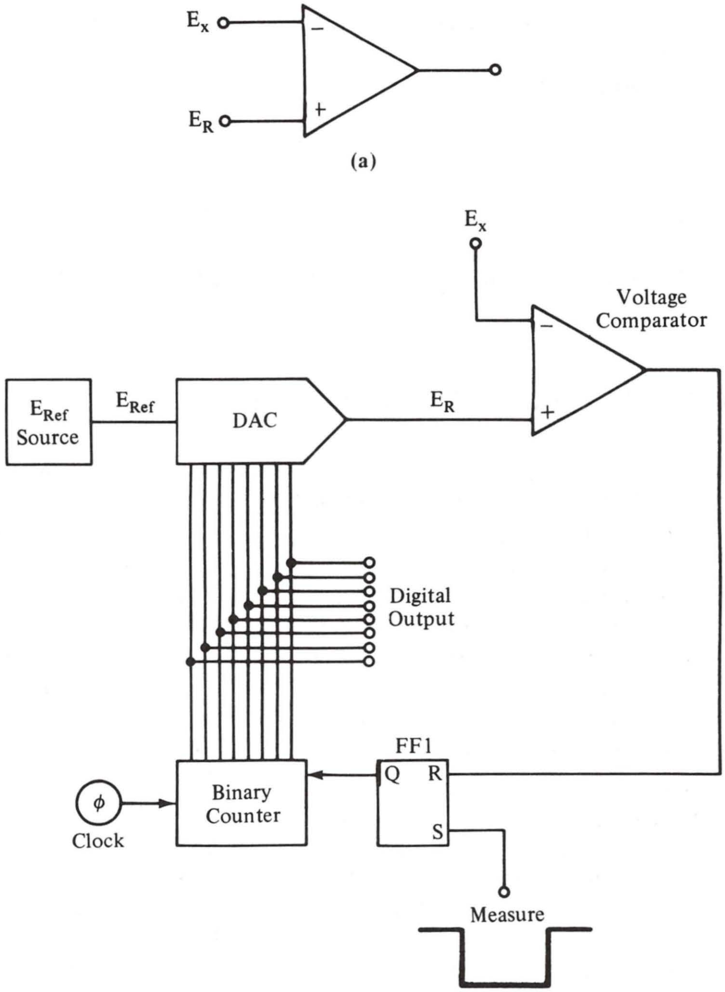
The unknown potential E_x is applied to one input terminal of the comparator, while a step-variable reference voltage E_r is applied to the other input.

Potential E_r is created by a precision dc voltage source and a *digital-to-analog converter* (DAC), as shown in Fig. 4-5(b). The DAC is a special circuit (see Chapter 14) that produces an output voltage that is the *product* of reference voltage E_{ref} and a fractional binary number applied to the respective digital inputs. The transfer equation for a DAC with an n-bit digital input word is

$$E_r = E_{ref} \times \frac{\text{(binary word)}}{2^n} \tag{4-3}$$

Example 4-3

An eight-bit DAC is connected to a 1.000-volt reference source. Calculate the full-scale value of E_r. (*Hint:* E_r is at full scale when the digital input word is 11111111.)



(a)

(b)

Figure 4-5. (a) Voltage comparator. (b) Simplified automatic comparison circuit.

SOLUTION

$(11111111_2 = 255_{10})$

$$E_r = E_{ref} \times (255/256) \tag{4-3}$$

$$E_r = (1.00) \times (255/256)$$

$$E_r = \mathbf{0.996\ V}$$

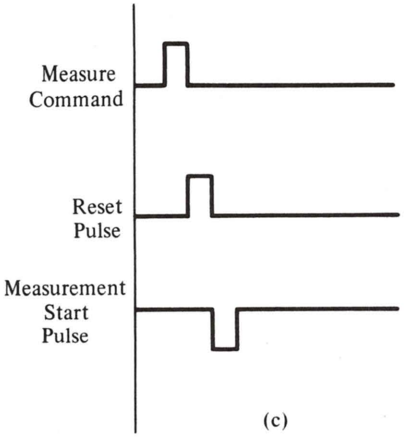


Figure 4-5. ***(cont.)*** (c) Timing scheme for (b).

Each count increments the DAC output by 3.9 mV, so the closest resolution of an 8-bit DAC connected to a 1.00-volt reference source is just under 4 mV. If Q represents the value of each DAC output voltage increment, then there is an uncertainty of $\pm Q/2$, or in this case, approximately 2 mV.

The operation of the electronic potentiometer is as follows:

1. Unknown voltage E_x is applied to the comparator.
2. The *measure* command causes the control logic section to generate a *reset* pulse to clear the binary counter to 00000000, followed by a *measure start* pulse [Fig. 4-5(c)].
3. The *measure start* pulse sets the Q output of R-S flip-flop FF1 HIGH, which turns on the counter.
4. As the counter increments once for each clock pulse, the DAC output E_r begins to rise.
5. The output of the comparator remains HIGH as long as E_x is greater than E_r, but drops to LOW when E_x equals E_r. When this occurs FF1 is reset, so its Q output goes LOW, thereby inhibiting the counter.
6. The value of E_x is determined by substituting E_x for E_r in Eq. (4-3).

The electronic potentiometer is used in digital instruments, A/D converters, and computer-controlled instrumentation processes and control systems.

4-9 SUMMARY

1. A *precision potentiometer* is used to create a precision voltage source.
2. The potentiometer may be used in either *null* or *equal-deflection* modes.
3. A digitial-to-analog converter (DAC) can be used to make an automated potentiometer in comparison measurements.

4-10 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread the appropriate sections.

4-11 QUESTIONS

1. Describe the construction of a slide-wire potentiometer.
2. Discuss why a slide-wire potentiometer produces a stable output voltage even when the operating temperature changes.
3. Describe in your own words how to measure an unknown potential E_x by using a Weston cell and a slide-wire potentiometer.
4. Describe a Weston cell; use a diagram if you need it.
5. Describe the *long-term* accuracy of an *unsaturated* Weston cell.
6. What temperature environment is required to maintain the accuracy of the output voltage calibration of a fresh Weston cell?
7. Describe the operation of an automated electronic potentiometer. What circuit elements are required?
8. Describe the principal differences between *null* and *equal-deflection* measurement techniques.

4-12 PROBLEMS

1. A potentiometer is set at the ¾ point such that $R_b = 0.75R$ [Fig. 4-1(b)]. Find E_0 if $E = 12.6$ volts dc.
2. A precision potentiometer is mechanically linked to a micrometer dial that reads from 00000 to 99999. Find E_0 if $E = 9$ volts dc and the dial is set to 63247.
3. A 10-bit DAC is connected to a 2.560-volt dc reference. Find the output voltage if the digital word at the input is 1011111111_2 (i.e., 767_{10}).

chapter 5

DIGITAL INSTRUMENTS

5-1 OBJECTIVES

1. To learn the operation of digital counting circuits.
2. To learn the stages and circuits needed to produce totalizer, frequency, and period counters.
3. To learn what factors in counter design affect accuracy.
4. To learn how to deal with signal-related problems that can cause count errors.

5-2 SELF-EVALUATION

Before studying the material in this chapter try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. How can a four-bit (i.e., modulus 16) binary counter be made into a decimal (i.e., modulus 10) counter?
2. What is the function of a *latch?*
3. How can amplitude modulation of a signal affect counting?
4. Some signals will not count because they do not cross ____________ of the hysteresis window.
5. Name three factors that affect the accuracy of a counter time base.

5-3 WHAT IS "DIGITAL"?

Most electronic instruments of recent design and manufacture are said to be "digital." But what *is* a digital instrument? What can a digital instrument do that an analog instrument cannot? Does being "digital" confer any advantage to the user?

Some people consider an instrument to be digital if the readout devices are digital displays. Such a readout will present the output in the form of lighted digits, such as those found on a calculator or electronic wristwatch. But many such instruments use ordinary analog electronic circuits, and then then use an ordinary voltmeter or current meter as the readout device. In medical electronics, for example, a transducer is used to convert blood pressure to an analogous dc voltage that is proportional to the pressure. A dc voltmeter, with the dial scale calibrated in units of pressure (i.e., mm Hg or Torr) is connected to the output of the pressure amplifier to serve as the readout.

Some manufacturers replace the analog voltmeter with a digital voltmeter (Chapter 7), but the remainder of the instrument remains unchanged. Is such an instrument "digital"? Not really, because only the readout device is a digital circuit. Such an instrument is not usually made any more accurate because of the digital readout, but it may give an indication that is less ambiguous, and therefore is easier to interpret. In that respect, the digital output meter may prove advantageous, if not more accurate.

Such digital instruments can, however, prove misleading if too many digits are used in the display. For example, if an instrument produces a dc output between 0 and 2 volts, with a resolution of 10 mV, then a digital voltmeter capable of measuring voltages of a lower level may tend to give the user more confidence in the end measurement than is justified. One popular DVM is capable of measuring voltages from 0 to 1.9999 volts, in 100-microvolt steps. Using this meter in an application where other factors would ordinarily indicate the use of a 0 to 1.99-volt instrument results in the user's gaining a false sense of precision—a fact that is, incidentally, not lost on some less scrupulous manufacturers. Not that they would lie, mind you, but some are guilty of creative specification writing.

A real digital instrument uses binary (i.e., base 2) logic elements such as gates, inverters, flip-flops, etc., to perform the measurement function. These circuits operate with only two voltage levels (see Fig. 5-1), and these represent the binary digits 0 and 1. In the medical monitor discussed above, it would be digital if the transducer passed the signal directly to an *analog-to-digital converter* with but minimal intermediate amplification or processing. All further signal processing functions would be performed in binary logic circuits, and the instrument would truly be "digital."

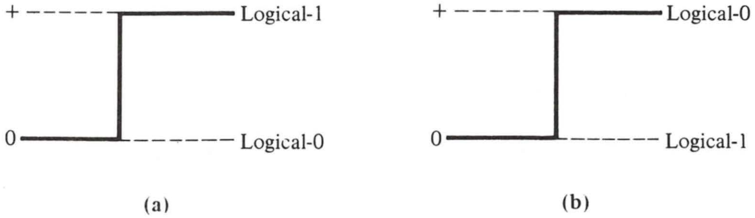


Figure 5-1. Digital instruments only recognize two voltage levels: (a) positive logic, (b) negative logic.

Although being digital does not automatically ensure improved accuracy or resolution, the use of certain digital design techniques often does improve the instrument. Do not, however, be misled into believing that an array of pretty digital readout devices automatically confers quality on the instrument.

5-4 BINARY COUNTERS

It is assumed that the student is familiar with basic digital logic circuits and the various digital integrated circuits that are available.

Be sure that you understand binary arithmetic, inverters, the various types of gates, and the different types of flip-flop used in digital electronics.

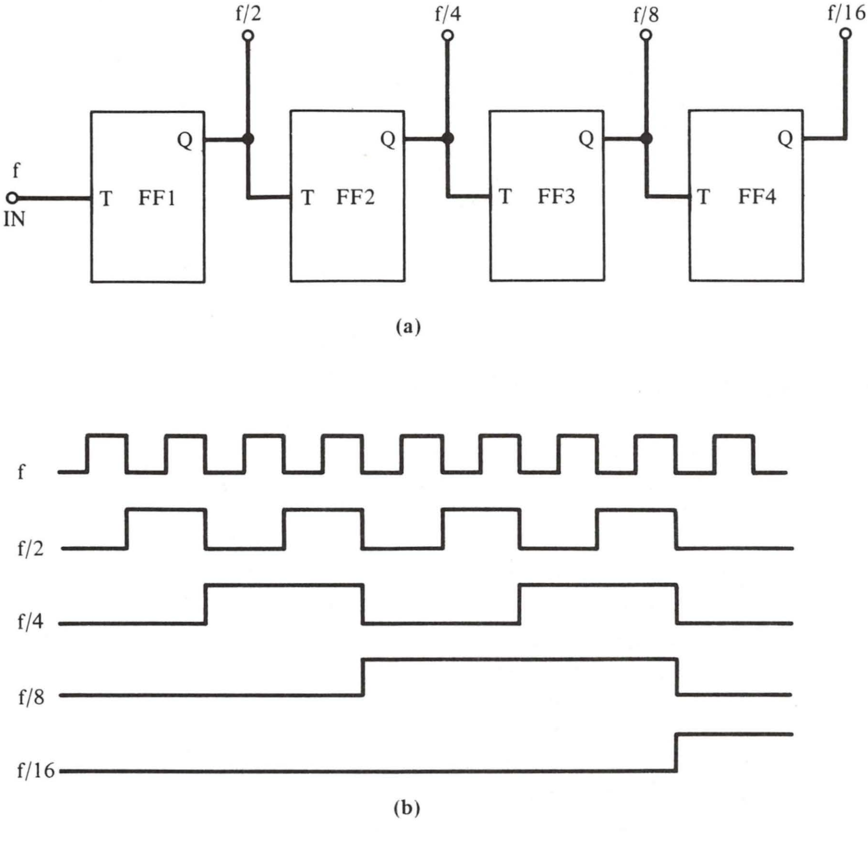


Figure 5-2. (a) Cascading flip-flops create a counter. (b) Logic states at outputs in (a).

A *binary counter* consists of a chain of clocked flip-flops [Fig 5-2(a)]; each stage is designed to count by two. A *decoder* is used to indicate the state of the count by examining the unique *Q* and not-*Q* states at the output of each stage.

A four-bit counter will count from 0000_2 to 1111_2 (that is, 0_{10} to 15_{10}), while a *binary coded decimal* (BCD) counter only counts from 0000_2 to 1001_2 (0_{10} to 9_{10}). Table 5-1 shows the BCD coding, which is a subset of four-bit binary coding.

Table 5-1 BCD CODES

Decimal	*8421 BCD*
0	0000
1	0001
2	0010
3	0011
4	0100
5	0101
6	0110
7	0111
8	1000
9	1001
10	0001 0000
11	0001 0001
12	0001 0010
13	0001 0110
14	0001 0100
15	0001 0101
16	0001 0110

5-5 DECIMAL COUNTING UNITS (DCU)

The heart of any electronic counter is a circuit called the *decimal counting unit,* or DCU. All DCUs consist of a *decade counter* and a *decoder/display-driver,* and most also include a *data latch.*

An example of a DCU is shown in Fig. 5-3. This circuit has been widely used, and is constructed of TTL integrated circuit logic elements. The decade counter is the type 7490 TTL IC, and uses a chain of flip-flops not unlike Fig. 5-2(a). But the 7490 also contains some internal gates that sense the tenth state (i.e., "9"), and resets the counter to 0000 if another input pulse is received.

Very few designers now use individual flip-flops to form decade counter circuits, because there are several complete IC decade counters available in both TTL and CMOS lines. In fact, there are several MSI and LSI integrated circuits available that contain *all* circuitry needed to form a complete electronic counter of several decades.

The four output lines from the decade counter in Fig. 5-3 are coded in the 8421 BCD format, and change state with every pulse. There are also two other

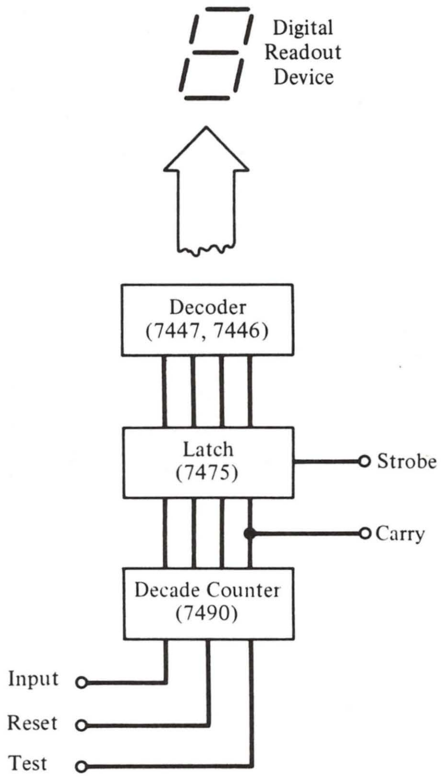


Figure 5-3. Decade or decimal counter.

lines: *input* and *reset*. The input line accepts the pulses being counted, while the reset line will reset the counter to 0000 when brought HIGH. For normal counting the line is kept LOW, and to clear the counter it is brought momentarily HIGH.

The decoder, in this case, is a 7446 or 7447 TTL IC, and is used to convert the four-bit BCD code to the code needed to correctly drive the readout device. The 7490 outputs can be used to drive the decoder directly, but this results in a difficult-to-read "rolling" display. The rolling effect is created because the display will change with the count. For example, if the count is to be "8," then the display will read "0-1-2-3-4-5-6-7-8" in sequence as the count accumulates to the "8" state.

The rolling effect can be eliminated by using a quad latch circuit (such as the TTL 7475). The latch is a bank of four Type-D flip-flops connected so that their *clock* terminals are tied together to become a *strobe* line. If the strobe line is high, then data appearing on the input lines are transferred to the output lines. But if the strobe line is low, then the output lines retain the last data present at the input when the strobe was high.

In normal practice, the strobe line is held low while the count is being accumulated, and at the end of the count period the strobe line is brought high momentarily, just long enough to transfer the count data from the input lines to the output lines. These data will then be held on the output lines until another strobe pulse updates the output to reflect the result of the next count.

5-6 DISPLAY DEVICES

Early electronic counters used a column of ten incandescent lamps (Fig. 5-4) for each digit. The decoder was a circuit that examined the BCD data lines, and determined which decimal digit was being represented. The decoder would then ground one of ten lines to light up the lamp that corresponded to that decimal digit. In this respect, the decoder is said to behave like a single-pole, ten-position switch that has its wiper grounded (see switch in Fig. 5-4). In the example shown, decoder output line 3 is grounded, so lamp 3 is lighted.

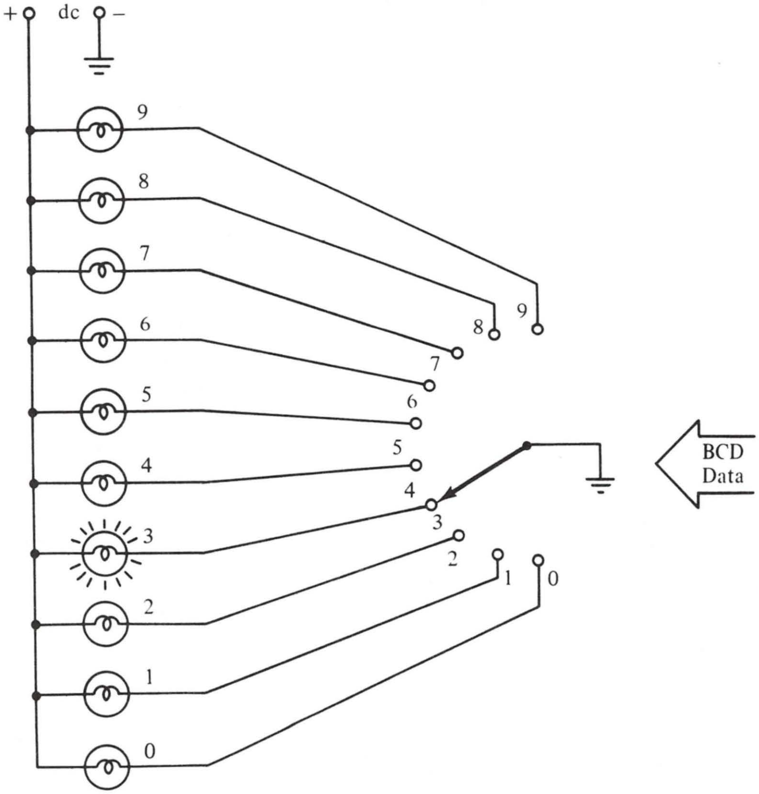
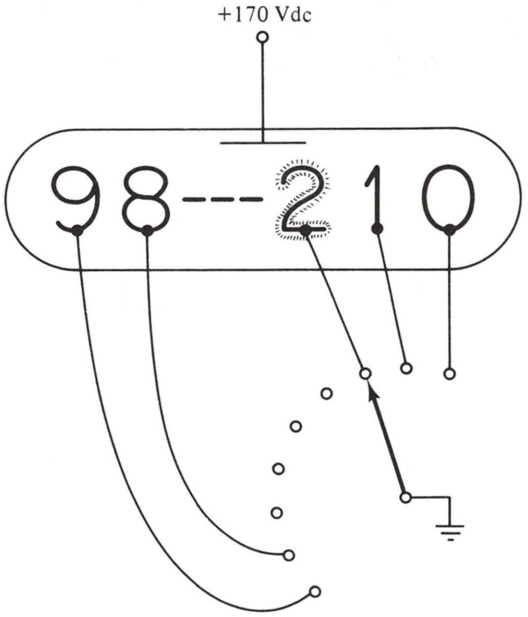


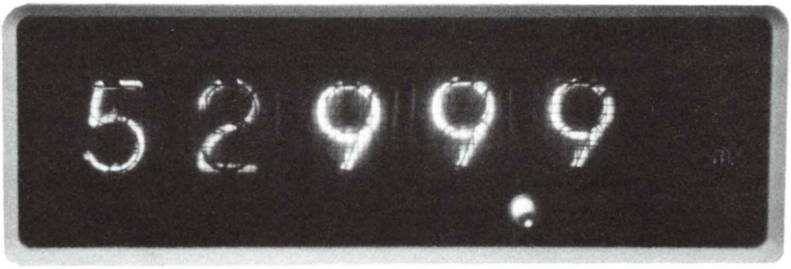
Figure 5-4. Principle behind simple BCD-to-decimal decoder.

The ten-lamp column display was used because it was very low-cost compared with the other alternatives then available. One of those alternatives was the Burroughs *Nixie®* tube display, shown in Fig. 5-5.

The Nixie® tube is a neon gas discharge device similar to a glow lamp such as the NE-2 or NE-51. The anode is a positive electrode common to all of the digits. But there are ten separate cathodes—one for each digit. The cathodes take the form of thin wire elements that are formed into the shape of digits 0 through 9.



(a)



(b)

Figure 5-5. (a) Nixie® tube operation. (b) Nixie® tubes.

When one of the digits is grounded, the neon gas particles close to its surface are ionized and give off light in the shape of the digit represented. All ten cathodes are arranged to that they face the same viewing plane.

The Nixie tube requires the same type of 1-of-10 decoder as did the lamp column, but it must be able to withstand the high voltages (e.g., 170 V dc) required to ionize the tube's gas.

The most common digital display in modern equipment is the seven-segment readout shown in Fig. 5-6. This type of display consists of seven illuminated bars arranged in a figure-eight pattern. It is possible to form all ten decimal digits by lighting appropriate bars [Fig. 5-6(b)]. A BCD-to-seven segment decoder will ground those terminals *a* through *g* required to form the decimal digit represented in BCD form at the decoder input.

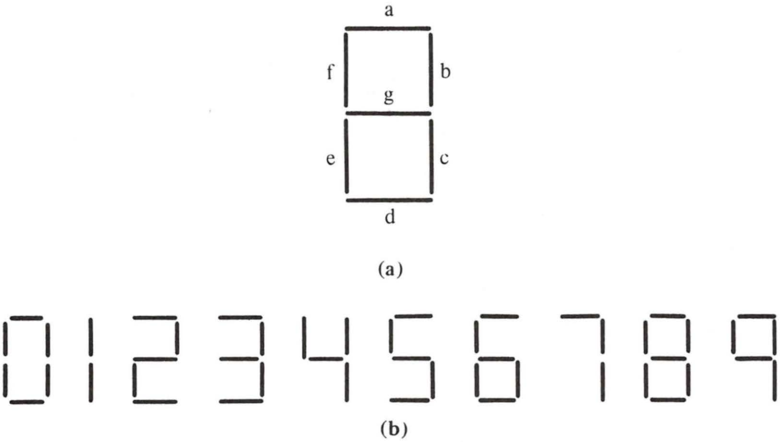


Figure 5-6. Seven-segment decimal readout.

Seven-segment readouts are available in a variety of different forms, using several different light sources. The RCA Numitron, for example, uses incandescent wire filaments to form the segments. This type of readout is constructed inside a modified seven- or nine-pin vacuum tube envelope.

Probably the most popular type as of this writing is the light-emitting diode (LED) display. In fact, the use of these displays is so common that many people erroneously refer to any seven-segment display as a "LED readout." In the LED type of display the segments are formed of (usually) red LEDs. Other forms of seven-segment readout are used, however, including *fluorescent* (greenish-blue) and gas discharge (yellow-orange). The *liquid crystal* display is slow to form, and is a dark gray or light gray in color, but has certain advantages. Among them is the ability to create any alphabetic, numeric, or graphic device on the display along with the seven-segment digital readout. Also, and this is a big advantage to designers of low-current-drain portable equipment, the liquid crystal display draws current only when the crystal is changing state. Once the number is formed, the device draws only a leakage current of a few microamperes. As a result, hearing aid batteries can be used to power CMOS calculators for 1500 hours.

One last type of display is the 5 × 7 dot matrix, shown in Fig. 5-7. This type of display is commonly used on athletic scoreboards and bank time-temperature billboards. It consists of 35 light sources arranged in five vertical columns, each containing seven light sources. This 5 × 7 matrix of light sources can be illuminated to form all numerical and alphabetic characters, and they look more natural than do seven-segment displays. The 5 × 7 dot matrix is also used in cathode ray tube (CRT) computer TV displays.

In the latter application, the points of light making up the matrix are unblanked points on a raster-scanned, but blanked, CRT screen. Each horizontal row of points of light represents one scan of the TV screen by the electron gun inside the CRT.

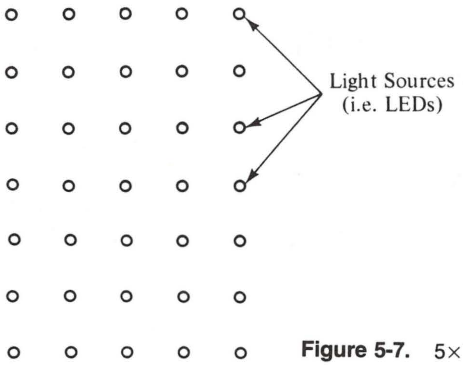


Figure 5-7. 5×7 matrix.

5-7 DECIMAL COUNTING ASSEMBLIES

The individual DCU can only count from 0 to 9, but if two DCUs are used in cascade the assembly can count from 0 to 99, and three DCUs can count from 0 to 999 (etc.).

Figure 5-8 shows a basic *decimal counting assembly* (DCA) consisting of three DCU elements. Three DCUs allow counting to 999, and one more DCU must be added to the assembly in cascade for each additional order of magnitude (i.e., for 0 to 9999 use four DCUs; for 0 to 99,999 use five DCUs, etc.).

The *strobe* and *reset* lines from all DCUs are tied to *common* strobe and

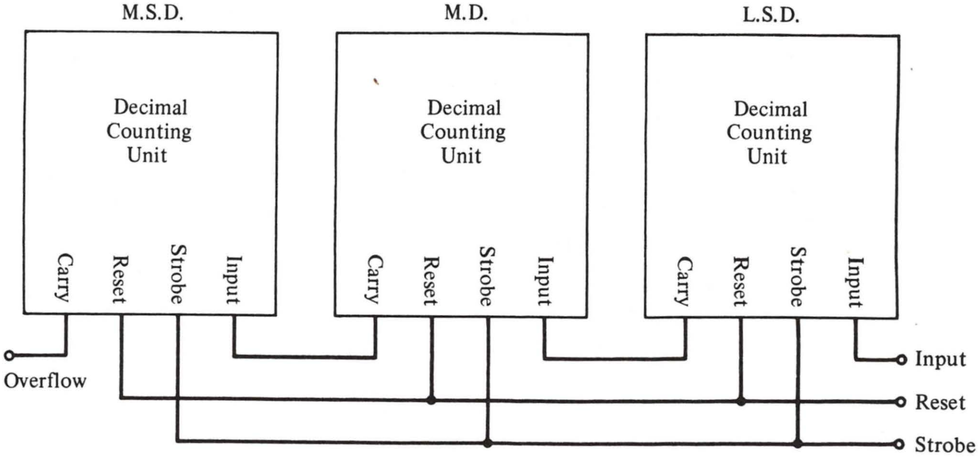


Figure 5-8. Decimal counting assembly.

reset lines that serve the entire DCA. All three DCUs in the assembly will be affected simultaneously by pulses applied to these lines.

The *carry* output of a DCU is the "D" data output line in any given DCU (i.e., the line weighted "8"), and it is used to drive the input of the next DCU in the cascade chain. Recall from your digital electronics studies that the J-K flip-flops used in decade counters change state only on a negative-going transition of the clock terminal. The BCD line that is weighted "8" is HIGH for both "8" and "9" states, but drops LOW when the decade counter overflows on the "10" count. It therefore toggles the input flip-flop of the next DCU *only* following the tenth count. This same tenth count also returns the counter to the "0" state.
counter to the "0" state.

The original signal to be counted, then, is applied to the input of the DCU that is in the least significant digit position in the DCA (LSD in Fig. 5-8). The *carry* output of LSD DCU is applied to the input of the next digit DCU, and finally, the output of that DCU is applied to the input of the most significant stage (i.e., MSD). The output of the MSD stage can then serve as a *counter overflow* indicator.

A counter constructed of only a basic DCA is called a *totalizer;* it accumulates input pulses as long as it is turned on and not reset. The DCA will simply totalize counts until it is either turned off or a reset pulse is applied. If the DCA overflows, the count starts back at zero, but continues to increment as pulses are received at the input. Of course, when using a totalizer counter one must be sure that the number of *events* (i.e., input pulses) expected is less than the maximum number that the counter will accept without overflowing.

5-8 FREQUENCY COUNTERS

Figure 5-9 shows the basic block diagram for a *frequency* counter. The sections include the *DCA, main gate, trigger, input amplifer, main gate flip-flop, time base,* and a *display clock.*

The DCA is a totalizer counter as shown in Fig. 5-8. The overflow stage is a flip-flop that is SET when the MSD carry output goes high. The overflow flip-flop turns on a lamp to make the operator aware of the overflow condition so that the data can be disregarded.

A frequency counter measures *events per unit of time* (EPUT), i.e., cycles per second, so the DCA must be turned on only for a given period of time (e.g., 0.1, 1, or 10 sec). The main gate, main gate flip-flop, and time base sections are used to allow input pulses into the DCA for the designated period of time.

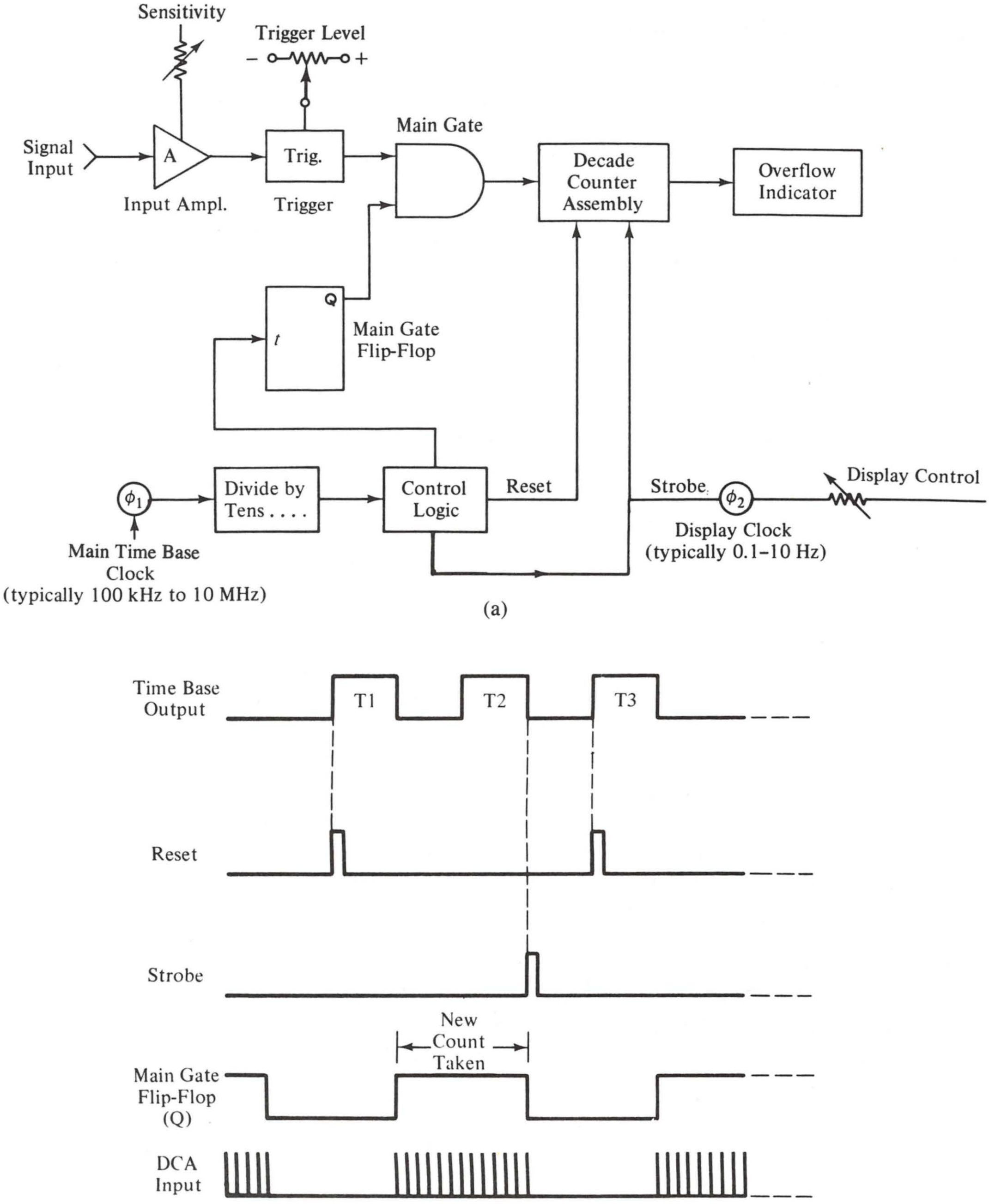
The time base section consists of a crystal oscillator that produces pulses at a precise rate such as 100 kHz, 1 mHz, 4 mHz, 10 mHz, etc. A chain of decade dividers such as the 7490 is used to reduce the crystal oscillator frequency to a lower frequency. The time base output frequency will be 10 Hz for 0.1 sec, 1 Hz for 1 sec, and 0.1 Hz for 10-second measuring periods.

The timing diagram for one complete interval of an EPUT counter is shown in Fig. 5-9(b). Pulses $t1$, $t2$, and $t3$ are output from the time base section. When pulse $t1$ goes high, the control logic section generates a short pulse to reset the DCA to zero. When $t1$ goes low again, the Q output of the J-K main gate flip-flop will go high. The main (AND) gate has one input tied to the Q output of the flip-flop, and the other input is tied to the signal being counted. As a result, the main gate passes input pulses to the DCA *only* when the Q terminal of the flip-flop is high.

The flip-flop remains set until the negative transition of $t2$ occurs. At that time the Q output of the flip-flop drops low, turning off the flow of pulses into the DCA, and causes the control logic section to generate a strobe pulse. This pulse tells the DCU latches to transfer data from the counter to the decoders. The display, then, shows only completed count cycles, and will hold the previous count until the end of the next interval. The trigger and input amplifier circuits will be discussed in Sec. 5-11.

The frequency counter of Fig. 5-9(a) counts *frequency,* i.e., events per unit of time, because the DCA is enabled only for a specific unit of time. The frequency of the input signal is the number of counts accumulated on the DCA divided by the time base period in seconds.

$$f_{(\mathrm{Hz})} = \frac{\text{counts on DCA}}{\text{time (sec)}} \tag{5-1}$$



(a)

(b)

Figure 5-9. (a) Block diagram of a frequency counter. (b) Timing diagram for a counter.

Example 5-1

What is the frequency of a signal in hertz if a five-digit DCA reads 08211 and the time base is set to 1 ms?

SOLUTION

$$f_{(\mathrm{Hz})} = \text{counts/time} \tag{5-1}$$

$$f_{(\mathrm{Hz})} = \frac{08{,}211 \text{ counts}}{0.001 \text{ sec}} = \mathbf{8{,}211{,}000 \text{ Hz}}$$

If the time base is 1 sec, the DCA reading will be the frequency in hertz, and if the time base is a decade multiple or submultiple of 1 Hz, then the frequency in hertz can be determined by adjusting the decimal point.

5-9 COUNTER DISPLAYS

The display used on a frequency counter will be one of those discussed earlier in the chapter, but there might be additional features added to make it easier for the operator to use, or to conserve battery power in the case of a portable instrument.

Ripple blanking is a feature of some DCA in which each DCU issues a signal that tells whether the count is zero or nonzero. If it is zero, then this ripple blanking output tells the ripple blanking input of the next least significant DCU. In this way, the counter is able to blank out, i.e., turn off, nonsignificant zeros to the left of the most significant number. In the case of Example 5-1, a five-stage DCA read 08211. If ripple blanking were used, the display would have read "8211"; the leading zero would have been suppressed. This makes the display less confusing and saves battery power in portable models.

Another tactic that saves both power and wiring effort is *display multiplexing*. This is a technique common in low-cost calculators and in much portable equipment. All of the seven segment lines (*a* through *g*) from all of the displays are tied together to form a seven-line *bus*. A timing control circuit will output the seven-segment code for only one digit at a time, along with a pulse that turns on only that digit. Each digit receives its code sequentially, so only one display is turned on at a time. But if the switching occurs rapidly enough, the operator's eye persistence causes the display to *appear* constant.

5-10 PERIOD COUNTERS

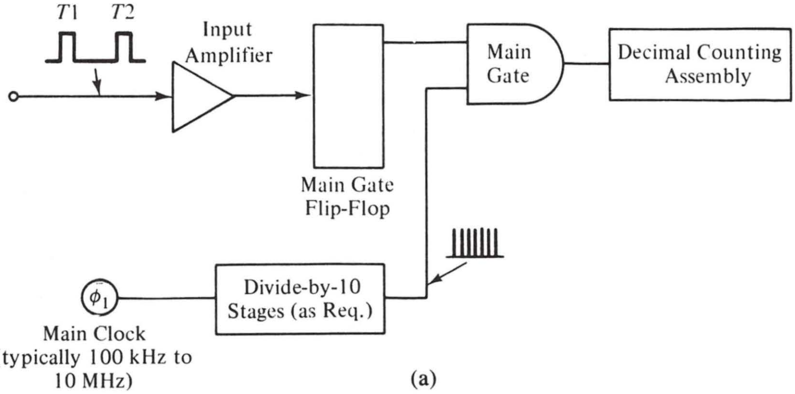
Period is defined as the time elapsing between identical features on successive cycles of a waveform, and can be calculated from the frequency; period is the reciprocal of frequency:

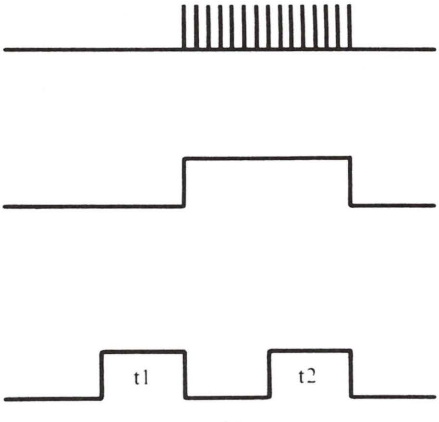
$$P_{(\mathrm{sec})} = \frac{1}{f_{(\mathrm{Hz})}} \tag{6-2}$$

A period counter can be made by reversing the roles of the input stage and time base; the input amplifier is connected to the main gate flip-flop and the time base is connected to the main gate [see Fig. 5-10(a)].

The timing diagram for a period counter is shown in Fig. 5-10(b). The flip-flop is set (i.e., Q goes high) on the negative-going transition of $t1$, and this allows the gate to pass pulses from the time base to the DCA.

The time base frequency determines the time interval represented by each pulse. For example, a time base frequency of 1000 Hz yields a period counter resolution of 1/1000 second, or 1 millisecond. Similarly, one 10-kHz time base


T1
T2
Input Amplifier
Main Gate
Decimal Counting Assembly
Main Gate Flip-Flop
ϕ_1
Divide-by-10 Stages (as Req.)
Main Clock (typically 100 kHz to 10 MHz)
(a)



(b)

Figure 5-10. (a) Block diagram of a period counter. (b) Timing diagram for period counter.

yields 0.1 ms resolution. The period is given by the count accumulated by the DCA and the time base frequency:

$$P_{(\mathrm{sec})} = \frac{(\mathrm{DCA\ count})}{f_{(\mathrm{Hz})}} \qquad (5\text{-}2)$$

Example 5-2

Find the period in seconds of a signal if the DCA count is 8026 and the time base frequency is 10 kHz.

SOLUTION

$$P_{(\mathrm{sec})} = (\mathrm{DCA\ count})/F_{\mathrm{Hz}} \qquad (5\text{-}3)$$

$$P_{(\mathrm{sec})} = \frac{8026\ \mathrm{events}}{\left[\dfrac{10^4\ \mathrm{events}}{\mathrm{sec}}\right]}$$

$$P_{(\mathrm{sec})} = \frac{8026\ \mathrm{sec}}{10^4} = \mathbf{0.8026\ sec}$$

The *resolution* is the smallest time interval that can be measured on the counter, and is defined as the reciprocal of the time base frequency.

5-11 TRIGGER CIRCUITS

The input signal will very rarely be the nice, clean, square waves required for proper operation of the digital logic circuit elements used to make a counter. The signals may also be too low in amplitude to operate the digital logic circuits, or may be too noisy. Remember, a TTL flip-flop needs to see fast rise and fall times (i.e., good square waves) and amplitudes greater than 1.4 to 2 volts, or they may not operate properly.

The input signal, then, is passed through two processing stages: an *amplifier* and a *trigger*. The amplifier is a wide-band voltage amplifier with enough gain to build up the minimum allowable signal (usually 25 to 100 mV) to a level great enough to drive the trigger stage (i.e., 500–1000 mV).

The trigger stage is a Schmitt trigger circuit with a built-in hysteresis. This type of circuit is used to clean up irregularly shaped signals by making them into square waves. Figure 5-11 shows the normal operation of a trigger circuit. The output snaps high when the input signal crosses the lower hysteresis limit, and

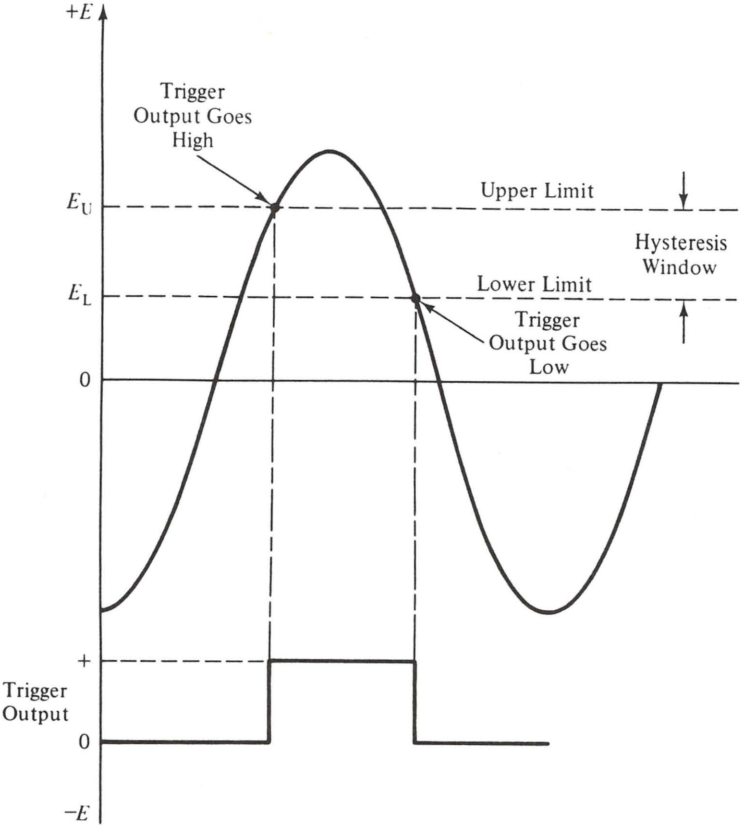


Figure 5-11. Normal operation of the trigger circuit.

high until the signal crosses the lower limit in a negative-going direction. The hysteresis window is the quantity $(E_u - E_L)$. Note that the trigger output possesses the shape and amplitude required by the digital circuits that it drives.

It is a fundamental rule that input signals must cross *both* hysteresis limits, or no count will be entered by the DCA. Figure 5-12(a) shows the required situation; the input sine wave crosses *both* limits. But in Fig. 5-12(b) the sine wave crosses only one of the window limits, so no count is registered on the DCA. Figure 5-13 shows some of the many types of signal that will *not* count when the hysteresis window is positioned symmetrical about zero volts.

Some counters have a *trigger level* control that allows the user to adjust the position of the window over a wide range. Other models use a three-position switch labeled "+," "preset," and "−." The switch allows the window to be placed in any of three locations (see Fig. 5-14). A continuously variable trigger level control allows positioning of the window *anywhere* within the range. Note

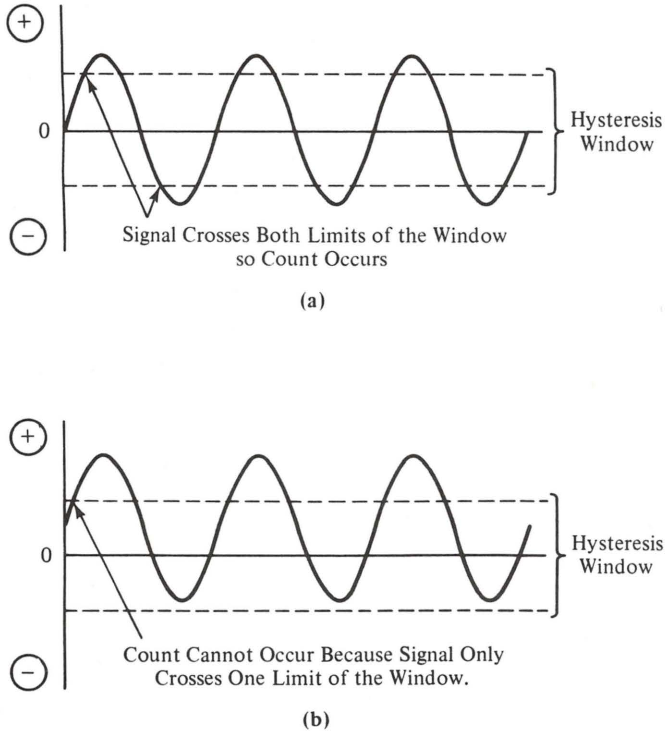

+
0
−
Hysteresis Window
Signal Crosses Both Limits of the Window so Count Occurs
(a)
+
0
−
Hysteresis Window
Count Cannot Occur Because Signal Only Crosses One Limit of the Window.
(b)

Figure 5-12. (a) Signal crosses both limits, so count occurs. (b) Only one limit is crossed by signal, so no count will occur.

that neither the continuously variable, nor the three-position-switch type of controls varies the width of the window (i.e., $E_{\mathrm{u}} - E_{\mathrm{L}}$), but does vary its position.

Some counters are equipped with a *trigger amplitude* control, which *does* allow the operator to vary the width of the hysteresis window.

5-12 COUNTER ERRORS

There are several factors that tend to reduce the accuracy of an electronic counter, and these can be grouped as *inherent errors* and *signal-related errors.* The inherent errors are a function of the quality, age, and history of the individual counter. Little can be done about these unless their source is a serious need for recalibration of the time base. Signal-related errors, on the other hand, are often correctable by proper manipulation of sensitivity, trigger level, and trigger amplitude controls.

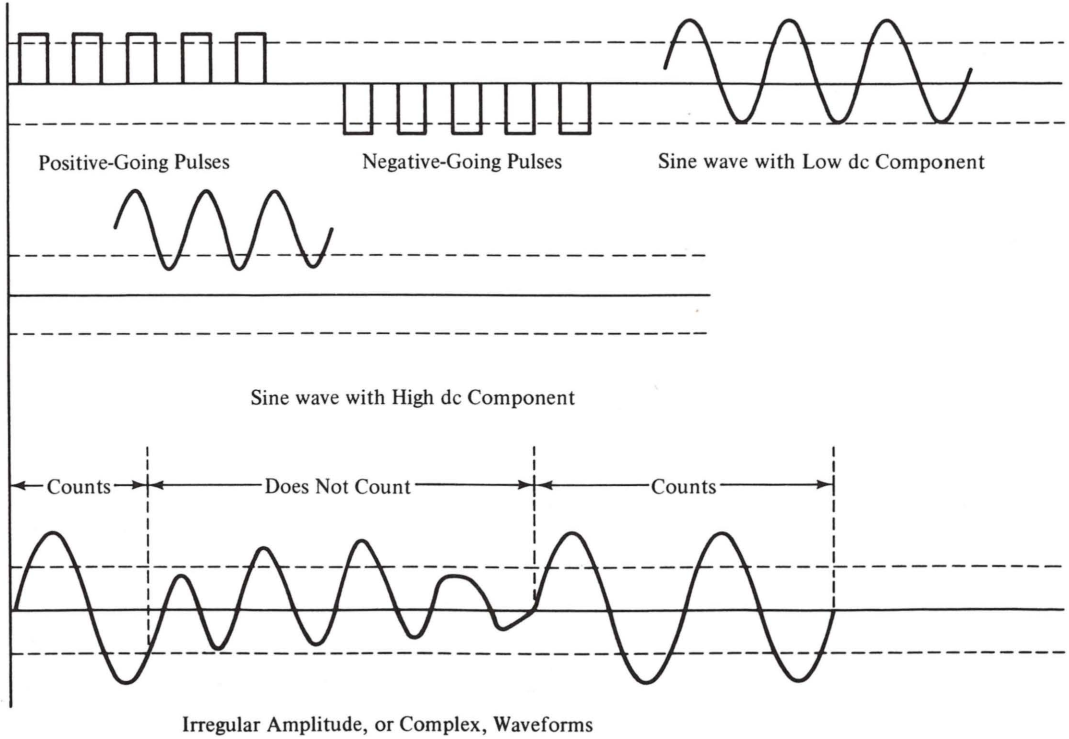

Positive-Going Pulses
Negative-Going Pulses
Sine wave with Low dc Component
Sine wave with High dc Component
Counts
Does Not Count
Counts
Irregular Amplitude, or Complex, Waveforms

Figure 5-13. Various signals that will not count.

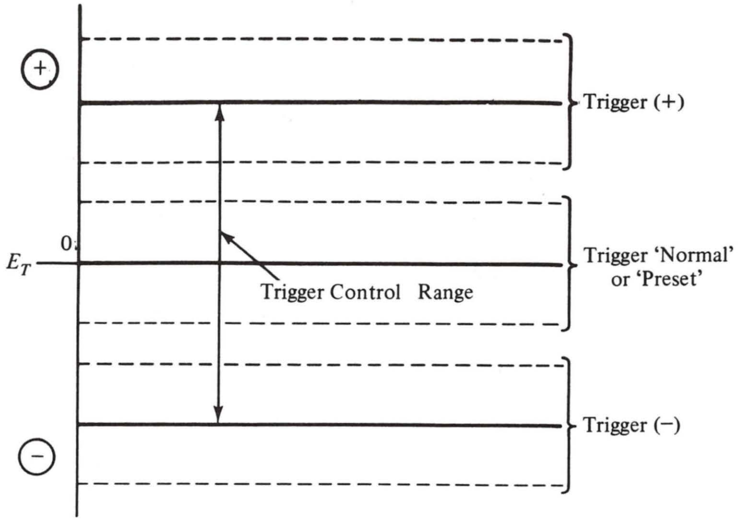


Figure 5-14. Range of the trigger control.

5-13 INHERENT ERRORS

There are two sources of inherent error in all frequency and period counters: time base error and a ±1 count ambiguity.

The time base error is expressed in terms of a percentage, or in parts per million. The error from time base inaccuracies is directly reflected in all measurements of frequency or period. For example, suppose a 1-mHz time base is off by 30 Hz, i.e., it is actually 1,000,030 Hz instead of 1,000,000 Hz. This is an error of 30 parts per million (30 ppm), which in percent is

$$\frac{1{,}000{,}030 - 1{,}000{,}000}{1{,}000{,}000} \times 100 = 0.003\%$$

The measurement error due to time base inaccuracy is constant regardless of the frequency being measured. That is to say, there will be a 0.003 percent error at 1 kHz and the same 0.003 percent error at the maximum frequency that the device will measure. For example, a 27-mHz signal would be measured with an error of

$$27\ \text{mHz} \times \frac{30\ \text{Hz}}{\text{mHz}} = 810\ \text{Hz}$$

This means that a counter reading 27,000,000 indicates that the actual frequency is 27 mHz ± 810 Hz. In other words, the actual frequency lies between

$$27{,}000{,}000 \text{ Hz} - 810 \text{ Hz} = 26{,}999{,}190 \text{ Hz}$$

and

$$27{,}000{,}000 \text{ Hz} + 810 \text{ Hz} = 27{,}000{,}810 \text{ Hz}$$

If the time base frequency is 30 Hz high, then the counter reading will be low, and if the time base frequency is low, then the counter reading will be high.

Total time base inaccuracy is the sum of several individual errors: *initial error, short term stability, long term stability, temperature change,* and *line voltage change.*

The initial error is the calibration error at the time the time base is initially adjusted at the factory, or at recalibration in a metrology laboratory. Different methods are used to measure the time base frequency, but in most cases the time base oscillator frequency is compared with standard frequency broadcasts of the *National Bureau of Standards* radio stations WWV, WWVB, or WWVH. Alternatively, it might be compared with the output of a cesium or rubidium beam atomic clock.

The *short term stability* is the time base oscillator frequency drift *per day. Long term stability* is the frequency drift *per month,* and is often designated the *aging rate.*

The temperature and line voltage stability specifications refer to the frequency change over the 0–50°C temperature range, and ±10 percent line voltage change, respectively.

There are four different classes of counter time base: *ac line, room temperature crystal oscillator, temperature compensated crystal oscillator* (TCXO) and *oven-controlled crystal oscillator.*

The use of the 60-hertz ac line as a counter time base is limited to the very cheapest models, and a few low-grade older units. Even low-cost units today have a crystal oscillator for the time base, and even though the crystal is operated at room temperature, it provides better accuracy than the 60-Hz power mains. Note that power companies will typically quote very high accuracy figures for their power plant's operating frequency, but these are frequency *averages* over a very long time. The short term accuracy, which is what concerns counter users, is terrible.

The TCXO is an encapsulated oscillator that is specifically compensated against temperature changes. The TCXO provides at least an order of magnitude better stability than room temperature oscillators. The TCXO is less expensive now than in the past, so even moderately priced counters now offer TCXO stability.

The oven-controlled crystal oscillator places the crystal, and in some cases the rest of the oscillator circuitry, inside an oven or thermal chamber. Thermostat ovens are considered an order of magnitude better than TCXO designs, while the *proportional control* type of oven is from one to two orders of magnitude better than TCXO.

Table 5-2 lists typical stability specifications for several models of counters by several different manufacturers. Note that the short term stability is given only for the oven type of time base. The TCXO and crystal oscillator must be operated for a full 24 hours before the stability reaches the specified level. At operating times less than 24 hours the stability is poorer. Some models use a separate regulated power supply for the TCXO that is not turned off by the main power switch. Rechargeable batteries are used in portable models for the same purpose, so the TCXO is not turned off while the counter is being transported between job sites.

Table 5-2 TYPICAL STABILITY SPECIFICATIONS

	XTAL	*TCXO*	*OVEN*
Long term aging (per mo.)	5×10^{-7}	2×10^{-7}	5×10^{-10}*
Short term aging (per day)	–	–	10^{-10}
±10% line voltage	10^{-7}	10^{-8}	10^{-9}
Temp. 0–50°C (ambient)	10^{-6}	10^{-7}	10^{-9}

*After 24-hour warmup.

The ± 1 count ambiguity is caused by the lack of synchronization between the input signal and time base. This is illustrated in Fig. 5-15. During period *t*1 ten pulses are gated into the DCA, while during *t*2 only nine pulses reach the DCA. On some subsequent count, it may be that nine pulses are gated into the DCA. One fundamental rule for all digital counter instruments is that there is an error of ±1 count of the least significant digit. In other words, a counter that reads, say, 10,000 Hz is measuring a frequency that lies between 9999 Hz and 10,001 Hz, i.e., 10kHz ± 1 Hz.

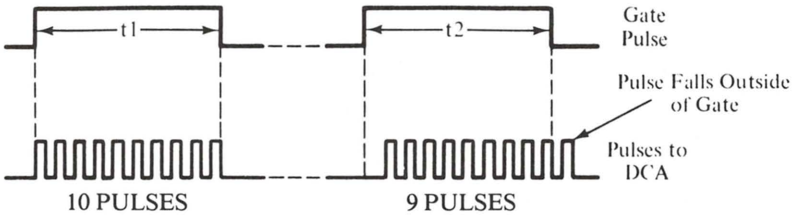


Figure 5-15. Mechanism for causing one-count ambiguity in counters.

The ±1 count ambiguity produces an error that is inversely proportional to the frequency being measured and the gate time.

$$\text{Error (\%)} = \frac{\pm 100}{fT} \qquad (5\text{-}4)$$

where f = the frequency being measured, in hertz (Hz)
T = the time the gate is open, in seconds (sec)

Example 5-3

Find the percentage of error due to ±1 count ambiguity at (a) 2 mHz and (b) 27 mHz for a gate time of 1 second.

SOLUTION

(a)
$$\text{Error (\%)} = \pm 100/fT \qquad (5\text{-}4)$$
$$\text{Error (\%)} = \pm 100/(2 \times 10^6 \text{ Hz})(1 \text{ sec})$$
$$\text{Error (\%)} = \pm \mathbf{0.00005\ percent}$$

(b)
$$\text{Error (\%)} = \pm 100/fT \qquad (5\text{-}4)$$
$$\text{Error (\%)} = \pm 100/(2.7 \times 10^7 \text{ Hz})(1 \text{ sec})$$
$$\text{Error (\%)} = \pm \mathbf{0.000004\ percent}$$

The error is ±1 count regardless of the frequency being measured, so the *percentage* of error decreases for higher frequencies. Compare (a) and (b) above.

5-14 SIGNAL-RELATED ERRORS

Poor signal quality can introduce errors that add to or subtract from the true count. Most of these errors result from hysteresis crossing errors or noise on the signal.

Trigger errors occur because the input signal crosses the hysteresis window limits *too many* or *too few* times. We saw in Figs. 5-12 and 5-13 that a signal will fail to increment the DCA if it does not cross *both* limits of the hysteresis window, causing too low a count.

Figure 5-16(a) shows how severe *ringing* on a signal can create extra, spurious counts of the DCA if the trigger level control is adjusted so that the ringing portions of the signal cross the limits, creating two *additional* ''input'' pulses, a two-count error.

The cure is to adjust the trigger level controls so that the ringing portions of the waveform fall outside the window limits [Fig. 5-16(b)].

The same problem exists on sine wave input waveforms (Fig. 5-17) that have a large amount of *harmonic distortion*. The cure is the same, however:

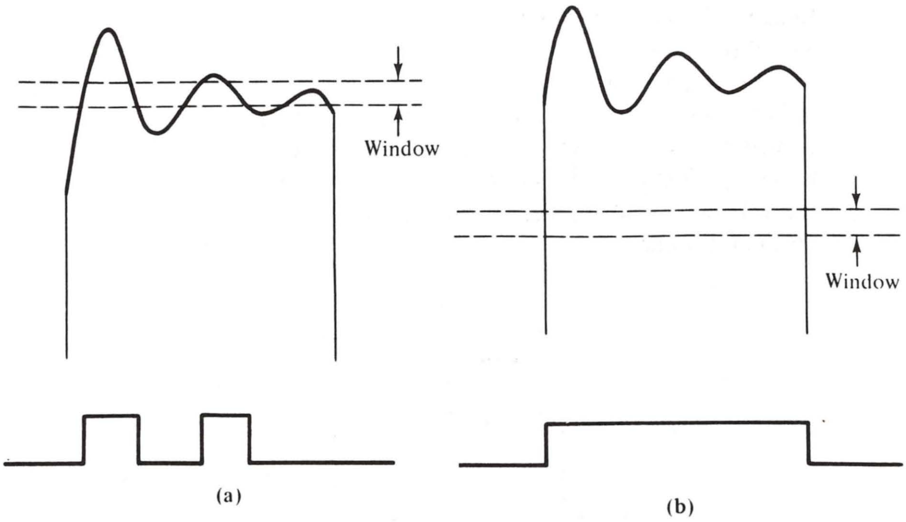


Figure 5-16. (a) Ringing causing false counts. (b) Proper adjustment of trigger to prevent problem in (a).

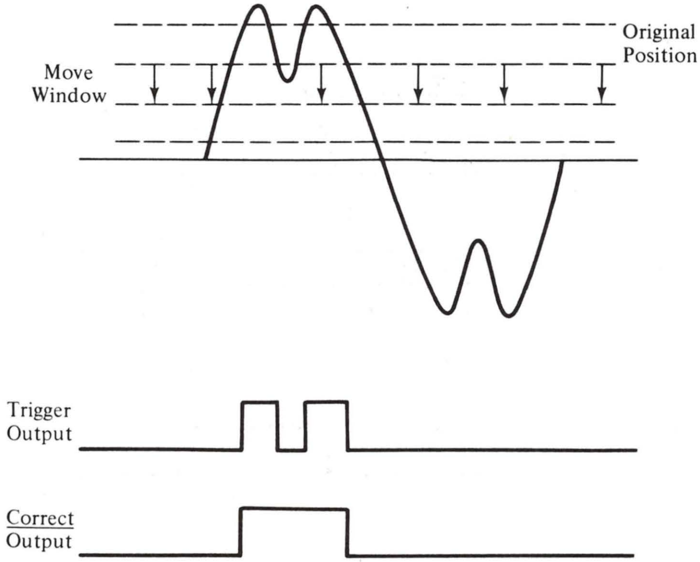
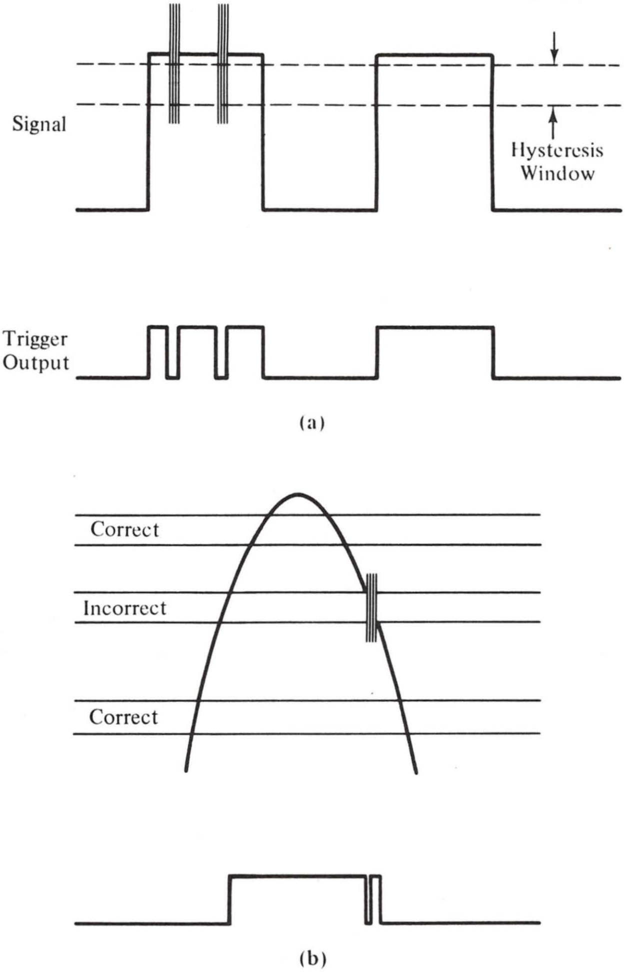


Figure 5-17. Second harmonic distortion causing false counts.

Readjust the trigger level control so that it is operating over a lower portion of the waveform.

Similarly, impulse noise riding on the signal can have an amplitude sufficient to cross both limits of the hysteresis window. An example of this phenomenon is shown in Fig. 5-18(a), in which a pulse in a symmetrical wave train is carrying impulse noise artifacts. In the case shown, the two noise bursts cross the window limits and thereby force the trigger output to create three pulses instead of just one.



(a)

(b)

Figure 5-18. (a) Oscillations on the waveform causing false counts. (b) Correct and incorrect settings of trigger for signals with noise or oscillations.

Once again the correction requires readjustment of the trigger level control to a point further down the waveform. In the case of a nonsquare wave, the noise may appear on the leading or trailing edges and still cause the problem. Figure 6-18(b) shows the proper and improper positions for the hysteresis window on such a waveform.

Note that filtering of the noise is not usually feasible, because of the bandwidth requirements of the input amplifier.

Figure 5-19 shows a type of error, caused by noise, that is particularly troublesome on period measurements. In this example noise rides on a signal that has a shallow slope, so creates a band of uncertainty around the signal. The trigger circuit should produce a high output when the signal crosses the upper limit, and drop low again when the signal crosses the lower limit. But noise impulses adding to, or subtracting from, the signal amplitude could provide premature, or delayed, trigger transitions. The correct duration of the trigger output pulse in Fig. 5-19 is $(t4 - t1)$, but under worst-case conditions the actual duration may be as much as $t5 - t_0$, and that amount represents a considerable error.

The solution for this problem is to cause the signal to slew through the hysteresis band as *rapidly* as possible. Two methods can be used to implement this solution. One is to narrow the window by adjusting the trigger amplitude control, whereas the other is to increase the waveform's slope by preamplification.

On some types of signal waveform it is sufficient to adjust the trigger level control so that the counter triggers on the *steepest* portion of the waveform. On sine waves, for example, this point occurs at zero crossings, but on other waveforms it may occur elsewhere on the signal.

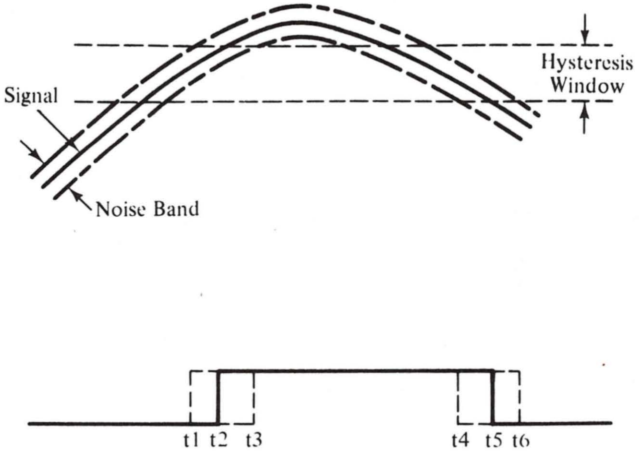
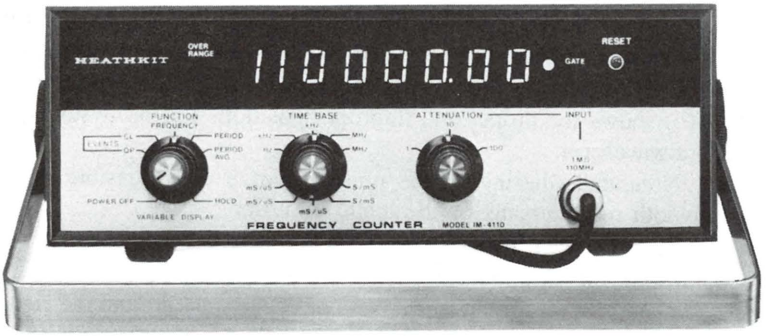
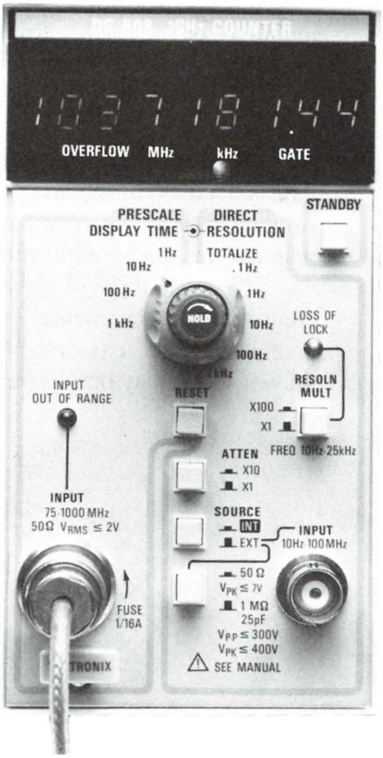


Figure 5-19. Counter errors due to hysteresis window and noisy signal.



(a)



(b)

Figure 5-20. (a) Typical independent frequency counter. (Courtesy of the Heath Co.) (b) Mainframe frequency counter (Courtesy of Tektronix, Inc.)

5-15 COMMERCIAL EQUIPMENT

Figure 5-20 shows two typical commercial frequency counters. The example in Fig. 5-20(a) is the Heath model 1M-4110. This model is a low-cost, independent, i.e., self-contained, counter that operates to 110 mHz with a stability of ±10 ppm, and an aging rate less than 10 ppm/year.

The Tektronix DC-508 frequency counter is shown in Fig. 5-20(b). With internal prescaling, this model is able to operate out to a frequency of 1 gHz (i.e., 1000 mHz). It features two inputs. A 50-ohm input is used in the 75–1000-mHz range, while a normal high impedance input is used in the 10 Hz to 100 mHz range. This model is part of Tektronix' TM-500 series, so it will fit the power supply main frames of that series. This feature means that the same counter can be used both in permanent bench applications and in luggage-portable configurations. The Tektronix DC-5010 programmable universal counter/timer is an example of an upscale modular instrument (see Fig. 5-20(c)). This instrument measures into the microwave region.

5-16 SUMMARY

1. Digital instruments use binary logic elements to perform circuit functions.
2. Frequency, period, and totalizing counters can be built from flip-flops, decoders, and digital display devices.

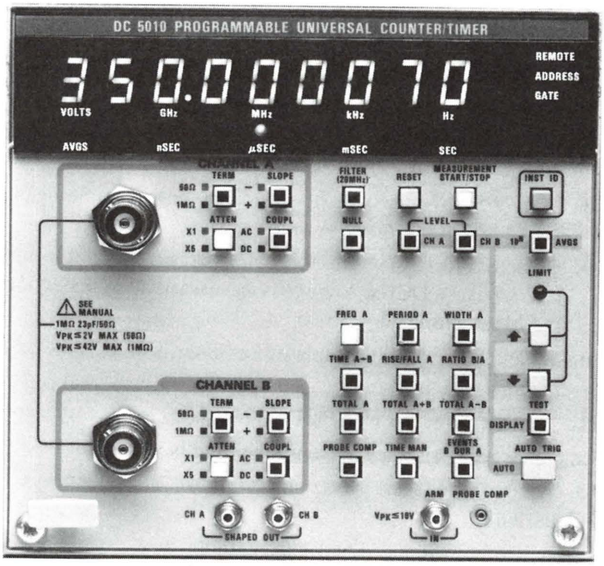


Figure 5-20. *(cont.)* (c) Tektronix DC-5010 programmable universal counter/timer. (Courtesy of Tektronix, Inc.)

3. Frequency is counted by opening a gate to allow input pulses into the DCA for a specified period of time.
4. Period is counted by opening the gate to admit time base pulses into the DCA between two successive cycles of the input signal.

5-17 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread the appropriate sections.

5-18 QUESTIONS

1. Does replacement of a taut-band voltmeter with a digital voltmeter as the readout device in an electronic instrument generally make it a *digital* instrument?
2. A 0 to 1 volt signal normally has a ±1 percent (that is, 10 mV) error. How many digits of a 0 to 999 mV digital voltmeter represent valid data?
3. Draw a block diagram for a simple *decimal counting unit* with *latch*.
4. In a latch circuit such as the 7475, data to the display are updated when the *strobe* line goes __________.
5. Name three forms of counter decimal display.
6. Which type of digital display device is used most frequently?
7. Discuss the relative advantages and disadvantages of the *Nixie*® tube and seven-segment readout.
8. Draw a 5 × 7 matrix display; indicate how it might display the decimal digit "8."
9. Draw a block diagram for a four-stage decimal counting assembly (DCA) using DCUs. Show any necessary interconnections and indicate the MSD and LSD.
10. Which line in a BCD output of a counter is used as the *carry?* What is its decimal weight?
11. Another name for an EPUT counter is __________ counter.
12. What type of stages make up the time base section?
13. What changes are necessary to make a frequency counter measure period?
14. Part of the input signal conditioning is a __________, which has upper and lower hysteresis window limits.
15. A *trigger level* control varies __________.
16. A *trigger amplitude* control varies __________.

17. (a) Define in your own words the following terms: short term stability, long term stability, ±10% power supply stability, and 0 to 50°C stability. (b) In which stage of a counter do these terms apply?
18. Rank the following in terms of long term stability beginning with the least stable: oven-controlled, TCXO, room temperature crystal oscillator, and 60-Hz mains.
19. Which is usually more stable—proportional controlled oven or thermostat-controlled oven?
20. Describe how noise on a signal can create an erroneous frequency count.
21. Describe how amplitude modulation of a signal could result in an erroneous frequency count.
22. Describe how noise on a signal can create trigger error when the period of a slow slew rate signal is measured.
23. What steps may be taken to alleviate the problem discussed in Question 22?

5-19 PROBLEMS

1. A DCA reads 12496 when the time base is set to 0.1 sec. What is the frequency of the applied signal?
2. A six-decade DCA is used to measure a 1256-Hz signal when the time base is set to 10 seconds. What reading would you expect on the display?
3. What reading would you expect to see on the DCA display in Problem 2 if the frequency is 1.567 mHz, and the time base is 10 microseconds?
4. What is the period of a 2.56-mHz signal?
5. What is the period of the signal if the DCA count is 12567 and the time base frequency is 10 kHz?
6. Find the period if the DCA count is 8067 and the time base period is 0.1 ms.
7. What DCA reading would you expect if the period of the input signal is 881 ms and the time base frequency is 1 kHz?
8. Find the percentage of error due to ±1 count ambiguity at (a) 100 kHz, (b) 1 mHz, (c) 2 mHz, (d) 20 mHz, and (e) 512 mHz.
9. A 0.005 percent error represents ____________ ppm.
10. A 10 ppm error represents ____________ percent.
11. Find the error in percent *and* ppm if a transmitter that is precisely tuned to a frequency of 20.005 mHz produces a reading of 20,004,682 Hz on a counter.

5-20 REFERENCES

1. Robert S. Stein, "Understanding and Using Electronic Counters," *Ham Radio* (Feb., 1978), pp. 10–20.
2. Hewlett-Packard applications note AN172: "The Fundamentals of Electronic Frequency Counters."
3. Joseph J. Carr, "Digital Frequency Counters for Servicing," Parts I and II, *Electronic Technician/Dealer* (Jan. and Feb., 1975).

chapter 6

MICROCOMPUTERS: AN INTRODUCTION

One of the first questions asked by a purchaser of a microcomputer is "what will it do?" This question is exasperating because it has so many *good* answers. Indeed, what *is* the role of a microcomputer? For that matter, what is a microcomputer? Microcomputers and microprocessors have long been a fact of life in process control, instrumentation, and other areas of engineering technology.

At one time, definitions were simple. As a freshman engineering student at Old Dominion University in Norfolk, Virginia, I was allowed to use an IBM 1601/1620 machine. *That* was a computer! There was no doubt in anyone's mind about that machine's identity; it took up an entire room on the second floor of the engineering school's building. But, today, an engineering student can sit at a small desk with an Apple II (complete with video CRT display, printer, and two disk drives) that has more computing power than that old 1601. Many engineering school students find that the cost of the typical small system is affordable and now they own their own computer. In fact, many engineering schools now require students to own a sophisticated microcomputer; they either bring one, buy one from the school, or rent one. The cost of the modern microcomputer is less than one-tenth what one of the lesser machines of only a decade ago cost, not counting the fact that 1973 dollars were bigger than today's dollars.

6-1 TYPES OF COMPUTERS

Before attempting to define the role of the microcomputer, let's first try to define the microcomputer. Terminology tends to become sloppy both from our own laziness and from the fact that once-genuine distinctions have become blurred as the state-of-the-art advances, terminology in the computer field is often overcome by events. For example, consider the terms *microcomputer* and *minicomputer*. Laziness tends to make some of us use these terms interchangeably, while modern single-chip computers (e.g., the Intel 8048 device) tend to make such usage seem almost reasonable. But, for our purposes, we require sharply focused meanings for

Adapted from *Elements of Microcomputer Interfacing* by J.J. Carr. Reston, VA: Reston Publishing Co., 1984, pp. 1–23.

these two terms and others: microprocessor, microcomputer, single-chip computer, single-board computer, minicomputer, and mainframe computer.

Microprocessor

The microprocessor is a large-scale-integration (LSI) integrated circuit (IC) that contains the *central processing unit* (CPU) of a programmable digital computer. The CPU section of a computer contains the *arithmetic logic unit* (ALU), which performs the basic computational and logical operations of the computer. The CPU also houses the *control logic* section (which performs housekeeping functions) and may or may not have several *registers* for the temporary storage of data. All CPU's have at least one temporary storage register called the *accumulator*, or *A register*. The principal attribute of a microprocessor is that it will execute instructions sequentially. These instructions are stored in coded binary form in an external *memory*.

Microcomputer

A microcomputer is a full-fledged programmable digital computer that is built around a microprocessor chip (i.e., integrated circuit); the microprocessor acts as the CPU for the computer. In addition to the microprocessor chip, the microcomputer typically will have some additional chips, the number varying from two to hundreds depending upon the design and application. These external chips may provide such functions as memory (both temporary and permanent) and input/output (I/O). The microcomputer may be as simple as a KIM-1 or as complex as a 30-board professional machine with all the electronic data-processing (EDP) options.

Single-Chip Computer

For several years we had no excuse for interchanging the terms *microprocessor* and *microcomputer;* a microprocessor was an LSI chip and a microcomputer was a computing machine. But the 8048 and similar devices began the process of dissolving the previously well defined boundaries because they were both LSI ICs and a computer. A typical single-chip computer may have a CPU section, two types of internal memory (temporary and long-term permanent storage), and at least two I/O ports. Some machines are even more complex.

The single-chip computer does, however, require some external components before it can do work. By definition, the microcomputer already has at least a minimum of components needed to perform a job.

Single-Board Computer

The single-board computer (SBC) is a programmable digital computer, complete with input and output peripherals, on a single circuit board. Popular examples are the KIM-1, SYM-1, AIM-65 and Z80 starter kit machines. The single-board computer might have either a microprocessor or a single-chip computer at its heart.

The peripherals on a single-board computer are usually of the most primitive kind (AIM-65 is a notable exception) consisting of seven-segment LED numerical displays and hexadecimal keypads reminiscent of those on a handheld calculator or Touchtone® telephone. The typical display is capable of showing only hexadecimal numeral characters, and even these are a little optimistic because of the form constraints of using seven-segment LED display devices. The Rockwell International AIM-65, on the other hand, uses a regular ASCII keyboard and a 20-character display made of 5 by 7 dot matrix LEDs. In addition, the AIM-65 has a built-in 20-column dot matrix thermal printer that uses paper similar to calculator and adding machine paper.

Most single-board computers have at least one interface connector that allows either expansion of the computer or interfacing into a system or instrument design. The manufacturers of SBCs, such as the KIM-1 and others, probably did not envision their wide application as a small-scale development system. These computers were primarily touted as trainers, that is, for use in teaching microcomputer technology. But for simple projects such computers also work well as a mini development system. More than a few SBC trainers have been used to develop a microcomputer-based product, only to wind up being specified as a "component" in the production version. In still other cases, the commercially available SBC was used as a component in prototype systems and then, in the production version, a special SBC (lower cost) was either bought or built.

Minicomputer

The minicomputer predates the microcomputer and was originally little more than a scaled-down version of larger data-processing machines. The Digital Equipment Corporation (DEC) PDP-8 and PDP-11 machines are examples. The minicomputer uses a variety of small-scale (SSI), medium-scale (MSI), and large-scale integration (LSI) chips.

Minicomputers have traditionally been more powerful than microcomputers. For example, they had longer length binary data words (12 to 32 bits instead of 4 or 8 bits found in microcomputers), and operated at faster speeds of 6 to 12 megahertz (mHz) instead of 1 to 3 mHz. But this is an area of fading distinctions. Digital Equipment Corporation, for example, offers the LSI-11 microcomputer that acts like a minicomputer. Similarly, 16-bit microcomputers are available, as are 6-mHz devices. It is sometimes difficult to draw the line of demarcation when a Z80-based microcomputer is in the same-sized cabinet as a minicomputer, and minicomputers can be bought in desk-top configurations.

Mainframe Computer

The large computer that comes to mind when most people think of computers is the *mainframe computer*. These computers are used in large-scale data-processing departments. Microcomputerists who have an elitist mentality sometimes call mainframe computers "dinosaurs." But, unlike their reptilian namesakes, these dinosaurs show no signs of extinction and are, in fact, an evolving species. The IBM 370 and CDC 6600 are examples of mainframe computers.

6-2 ADVANTAGES OF MICROCOMPUTERS

That microcomputers have certain advantages is attested by the fact that so many are sold. But what are the advantages, and how are they conferred?

The most obvious advantage of the microcomputer is *reduced size;* compared with dinosaurs, microcomputers are mere lizards! An 8-bit microcomputer with 64K bytes of memory can easily fit inside a tabletop cabinet. The Heath/Zenith H89 (Figure 6-1), for example, fits the complete computer (plus one optional disk drive) into the wasted space found inside their H19 video terminal cabinet. Another company packs a computer with 16K of random-access memory (RAM) inside a keyboard housing.

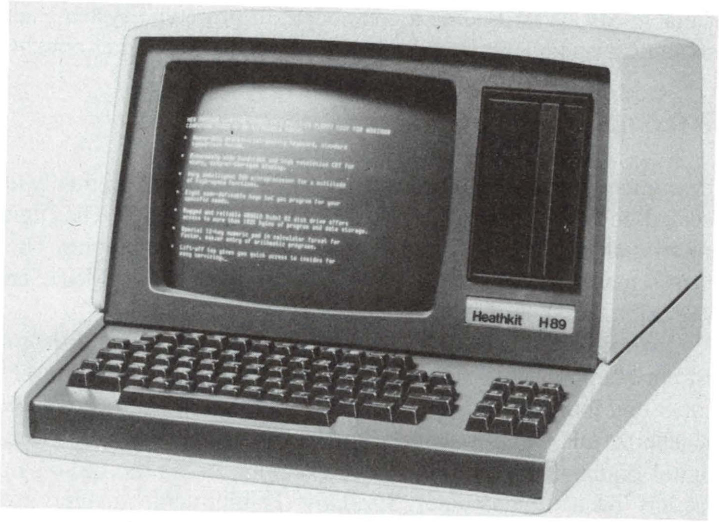


Figure 6-1. New Heathkit H89 All-In-One Computer features two Z80 processors, floppy disk storage, smart video terminal, heavy-duty keyboard, numeric keypad and 16K RAM, all in one compact unit. (Courtesy of Heath/Zenith)

The LSI microcomputer chip is generally more complex than a discrete-components circuit that does the same job. The interconnections between circuit elements, however, are much shorter (micrometers instead of millimeters). Input capacitances are thereby made lower. The metal-oxide semiconductor (MOS) technology used in most of these ICs produces very low current drain; hence the overall power consumption is reduced. A benefit of reduced power requirements is reduced heating. While a minicomputer may require a pair of 100 cubic feet per minute (cfm) blowers to keep the temperature within specifications, a microcomputer may be able to use a single 40-cfm muffin fan or no fan at all.

Another advantage of the LSI circuit is reduced component count. Although this advantage relates directly to reduced size, it also affects reliability. If the LSI IC is just as reliable as any other IC (and so it seems), then the overall reliability of the circuit is increased dramatically. Even if the chip reliability is lower than for lesser ICs, we still achieve superior reliability due to fewer interconnections on the printed circuit board, especially if IC sockets are used on the ICs. Some of the most invideous troubleshooting problems result from defective IC sockets.

6-3 MICROCOMPUTER INTERFACING

The design of any device or system in which a microcomputer or microprocessor is used is the art of *defining* the operation of the system or device, *selecting the components* for the device or system, *matching* and *integrating* these components (if necessary), and *constructing* the device or system. These activities are known collectively as *interfacing*.

But let's get down to a more basic level. Most readers of this book are technically minded people with some knowledge of electronics and computer technology. For most readers, therefore, interfacing consists of selecting and matching components and then connecting them into a circuit that does a specific job. These are the matters that are addressed in later chapters.

6-4 MICROCOMPUTERS IN INSTRUMENT AND SYSTEM DESIGN

Designers in the past used analog electronic circuits, electromechanical relays (which sometimes precipitate a maintenance nightmare), and other devices in order to design instruments, process controllers, and the like. These circuit techniques had their limitations and produced some irritating results; factors like thermal drift loomed large in some of these circuits. In addition, the design was cast in cement once the final circuit was worked out. Frequently, relatively subtle changes in a specification or requirement produced astonishing changes in the configuration of the instrument; analog circuits are not easily adapable to new situations in many cases. But, with the advent of the microcomputer, we gained the advantage of flexibility and solved some of the more vexing problems encountered in analog circuit design. The memory of the computer tells it what to do and that can be changed relatively easily. We can, for example, store program code in a *read-only memory* (ROM), which is an integrated-circuit memory. If a change is needed, the software can be modified and a new ROM installed. If the microcomputer was configured in an intelligent manner, it is possible to redesign only certain interface cards (or none at all) to make a new system configuration. An engineer of my acquaintance built an anode heat computer for medical x-ray machines. A microprocessor computed the heating of the anode as the x-ray tube operated and sounded a warning if the limit of safety was exceeded, thus saving the hospital the cost of a $10,000 x-ray tube. But different x-ray machines require different interfacing techniques, a problem that previously had meant a new circuit design for each machine. However, by intelligent engineering planning, the anode heat computer was built with a single interface card that married the ''universal'' portion of the instrument with each brand of x-ray machine. Thus, the company could configure the instrument uniquely for all customers at a minimum cost.

Another instrument that indicates the universality of the microcomputer is a certain cardiac output computer. This medical device is used by intensive-care physicians to determine the blood-pumping capability of the heart in liters per minute. A bolus of iced or room-temperature saline solution is injected into the patient at the ''input'' end of the right side of the heart (the heart contains two pumps, a right and left side, with the right side output feeding the left side input via the lungs). The temperature at the output end of the right side is monitored and the time integral of temperature determined. This integral, together with some constants, is used by the computer to calculate the cardiac output. These machines come in two versions, research and clinical. The researcher will take time to enter certain constants (that depend upon the catheter used to inject saline, temperature, and other factors) and will be more vigorous in following the correct procedure. But in the clinical setting, technique suffers owing to the need of caring for the patient, and the result is a perception of ''machine error,'' which is actually operator error. To combat this problem, the manufacturer offers two machines. The research instrument is equipped with front panel controls that allow the operator to select a wide range of options. The clinical model allows no options to the operator and is a ''plug and chug'' model. The interesting thing about these instruments is that they are *identical* on the inside. All that is different is the front panel and the position of an on-board switch. The manufacturer's program initially interrogates a switch to see if it is open or closed. If it is open, then it ''reads'' the keyboard to obtain the constants. If, on the other hand, it is closed, the program branches to a subprogram that assumes certain predetermined constants, which are loaded on the buyer's prescription at the time the instrument is delivered. The cost savings of using a single design for both instruments are substantial.

6-5 SOME REAL PRODUCTS

The physical machinery that one identifies as a real microcomputer may well depend upon your point of view. Certainly, the Heath H89 machine shown in Figure 6-1 is a microcomputer. It contains a relatively large array of on-board RAM-type memory and can accommodate at least one disk drive right in the same cabinet with the CRT video terminal. But the H89 might not be the computer for all applications. In some cases, other machines will be more appropriate. In this section we will examine some of the different machines that are commonly found on the market. Inclusion in this discussion connotes neither endorsement of the product nor that some other manufacturer's product is not just as good.

Heathkit ET6800

The Heathkit model ET6800 (Figure 6-2) is strictly a student trainer and is probably the lowest-cost computer on the market. It consists of a simple keypad-operated computer that is based on the Motorola MC6800 8-bit microprocessor chip. The display consists of a simple array of seven-segment light-emitting diode (LED) numerals, much like those used in calculators and other devices. The cardboard cabinet no doubt contributes immensely to the low cost of the machine.

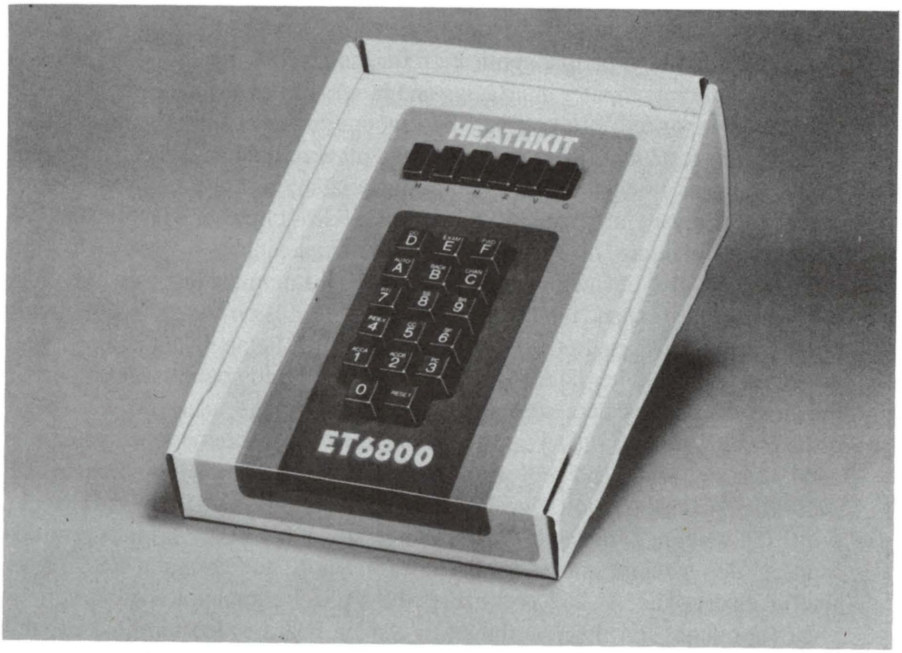


Figure 6-2. Small training "computer." (Courtesy of Heath/Zenith)

Heath originally designed the ET 6800 microcomputer as an aid to their continuing education course on microprocessors (EC-6800), and it is used to learn assembly language programming. The small computer contains a 1K-byte monitor program and permits access to 256 bytes of random-access memory (RAM).

The ET6800 is not a serious microcomputer for most users and is strictly used as a trainer. It is, however, a reasonable beginning for many students, especially in view of its low cost.

Synertek SYM-1

Several years ago, the original manufacturer of the 6502 microprocessor, MOS Technology, Inc., produced a small single-board computer that contained a hexadecimal keyboard and LED readouts. Originally conceived as a trainer, the KIM-1 microcomputer became something of a standard among single-board computers, and its bus is now sometimes referred to as the ''KIM-bus.'' The KIM inspired a large collection of magazine articles and books. For many of today's advanced computer sciences people, the little KIM-1 was their first introduction into the world of microcomputer technology.

Although the SYM-1 microcomputer shown in Figure 6-3 is based on the original KIM-1 machine, it extended the machine and provided more features than the original design. Synertek Systems Corporation of Santa Clara, California, is the manufacturer of the SYM-1 machine.

Probably the principal application for the SYM-1 is training engineers and students in microcomputer interfacing and programming technology, but applications have expanded into engineering laboratory work, prototyping of devices based on the 6502 microprocessor, instrumentation, and conducting both experiments and tests in engineering and scientific laboratories. The SYM-1 uses the identicial hardware interface bus as the earlier KIM-1 device, so it may be used in applications previously reserved to the KIM-1 machine.

The SYM-1 device has a 4K-byte on-board monitor program, 1K byte of on-board RAM (expandable to 4K bytes), and provision for up to 28K bytes of on-board ROM or programmable ROM (PROM). The applications port, which, again, is expandable, has 15 bidirectional, TTL-compatible I/O lines. The machine also offers data storage and program storage on audio cassettes (an ordinary cassette tape player, provided that it has both microphone and earphone jacks, can be used), and will accommodate a full-duplex teletypewriter (TTY) 20-milliampere (mA) loop. This last feature makes the SYM-1 compatible not just with TTY machines, but also with a wide variety of hard-copy printers now on the market. The machine includes one other I/O port, the common RS-232 serial interface port. The RS-232 port makes the SYM-1 compatible with a variety of video terminals and other peripherals. An on-board video terminal capability allows you to use either a TV monitor or, if a radio-frequency (RF) modulator is provided, a home TV receiver to receive output data (32-character line of video).

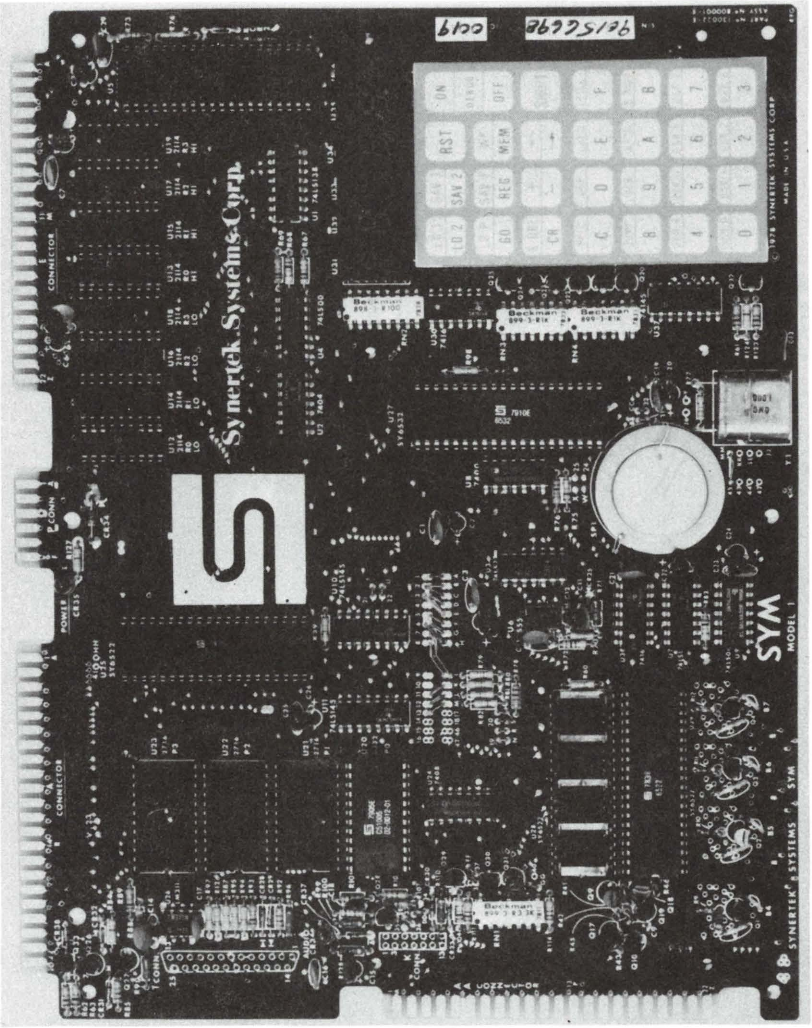


Figure 6-3. SYM-1 single-board computer. (Courtesy of Synertek Systems Corp.)

Ohio Scientific Superboard II

The microcomputer in Figure 6-4 solves some of the problems inherent in other single-board designs: inconvenient keyboard format, for one. This machine is also based on the same microprocessor (6502) as the KIM-1 and SYM-1 machines, although it does not use the KIM-1 bus. Programming and data entry are through a

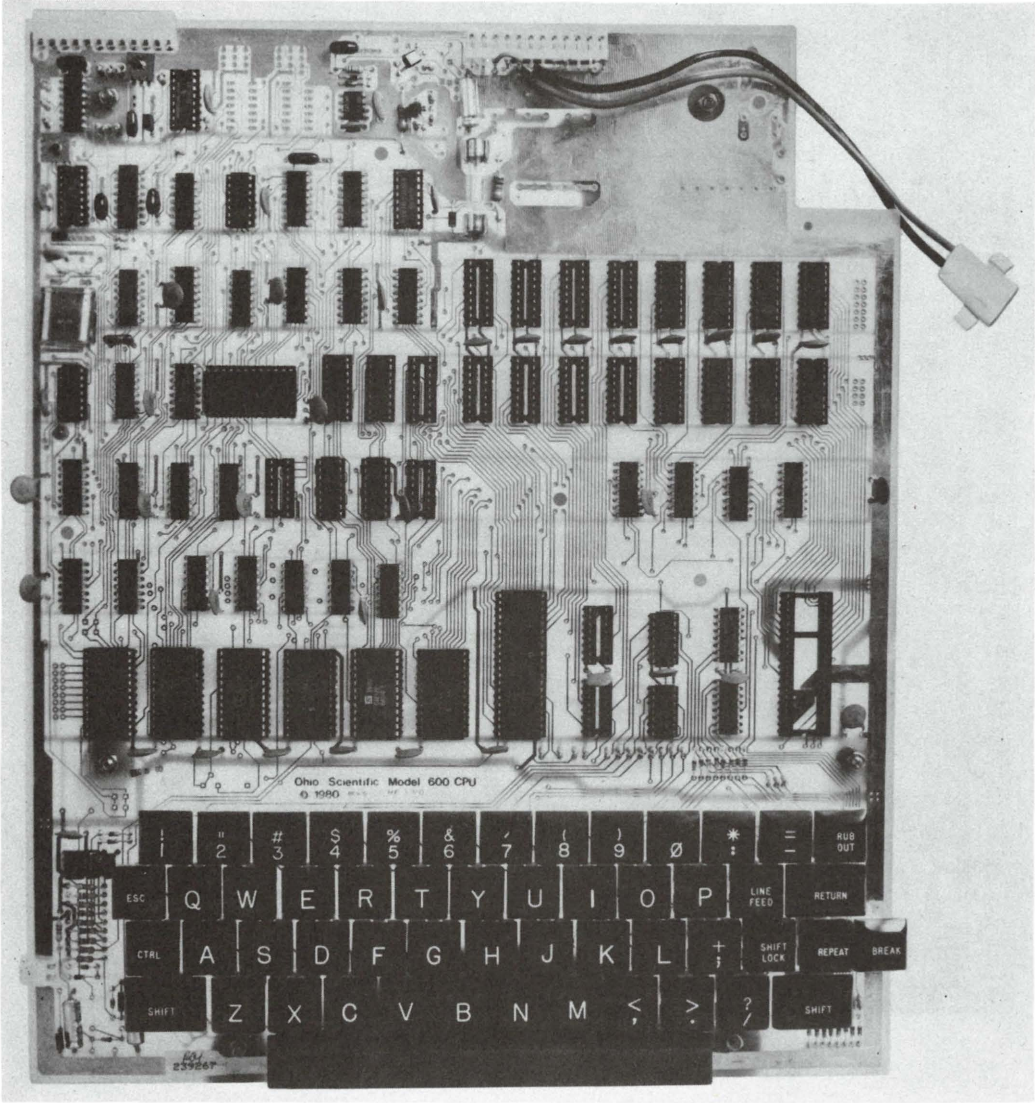


Figure 6-4. Ohio Scientific Superboard II. (Courtesy of Ohio Scientific)

full ASCII keyboard like those found on video CRT terminals and larger computers.

The Superboard II can interface with TTY, CRT video terminals, and other peripherals. It is probably one of the simplest of the ''advanced'' single-board computers and offers much that the lesser machines cannot (e.g., more memory and programming in BASIC).

Heathkit H-8

The model H-8 (Figure 6-5) was the Heath Company's first entry into the microcomputer market, so it has been around a long time. It is basically an 8080A-based machine, but a recent retrofit kit makes the H-8 into a Z80-based machine. The Z80 is still being used in great quantity, while the 8080A has faded somewhat in favor of more sophisticated chips.

The H-8 is a single-board computer, but it comes with a plastic cabinet to make it more attractive. Like other single-board computers, the H-8 uses a hexadecimal keyboard for programming and data entry and a series of LED seven-segment readouts for display.

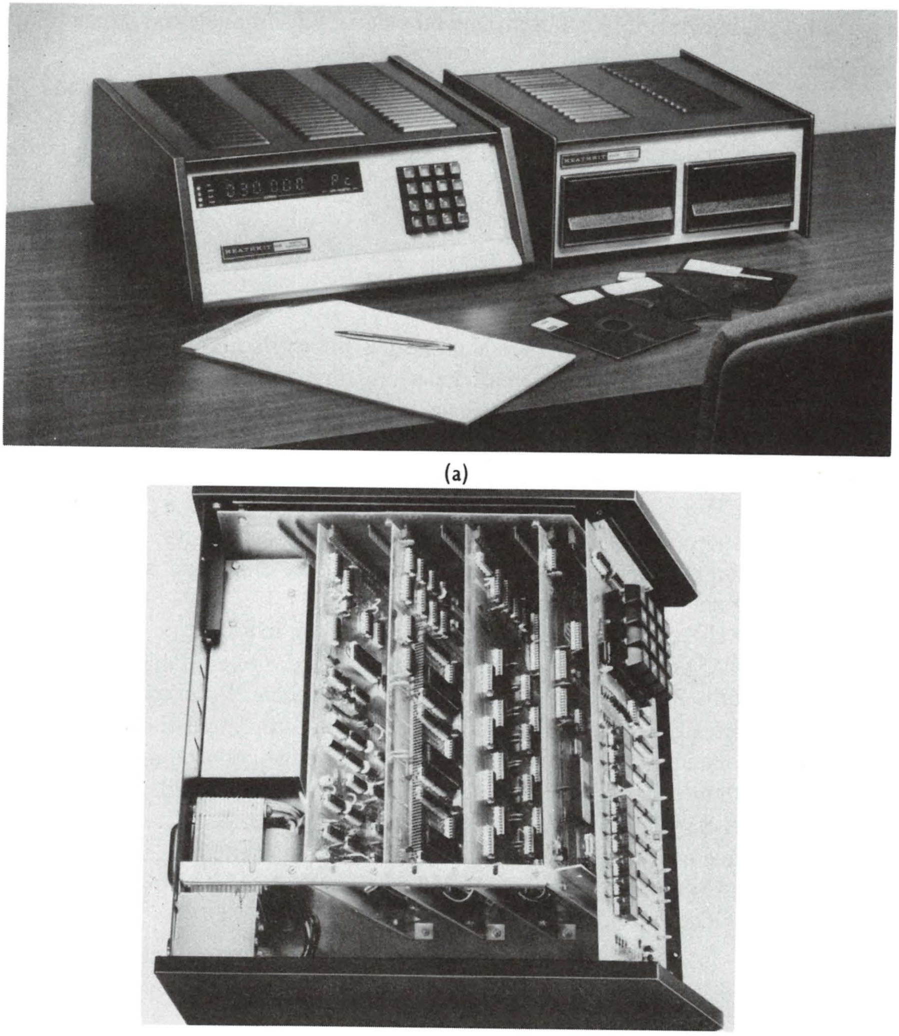
The H-8 contains a 1K-byte on-board monitor in ROM and comes with 16K bytes of RAM memory. The total memory package is expandable to the full 64K bytes that the 8080A and Z80 chips can accommodate. Extra memory and other features are added with the addition of extra plug-in printed circuit boards. The cabinet contains a total of seven extra PC card-edge connectors for expansion.

With the correct choice of interfacing and memory boards, it is possible to build an H-8 system that has 64K bytes of RAM, a pair of 5.25-inch floppy disk drives, three parallel I/O ports, a 1200-baud serial cassette port, and four RS-232 standard I/O ports (serial). The RS-232 ports make the H-8 compatible with most standard video terminals and other peripherals.

Apple II and III

The Apple II (and its later cousin, the Apple III), shown in Figure 6-6, have become the byword in personal computers, partially through an aggressive advertising campaign and partially because these computers make available a full-service microcomputer in a small package. It is quite possible to make a system that includes 48K bytes of memory, color TV graphics, color TV monitor, a teletypewriter, and two 5.25-inch disk drives and have it take up little more than a tabletop.

The Apple II comes with a plug-in BASIC, with a more extensive version of BASIC available as an option. It also has an assembly language and built-in disassembler capability. The ordinary Apple II is available with an audio cassette interface, although for any serious work it is recommended that at least one disk be acquired.



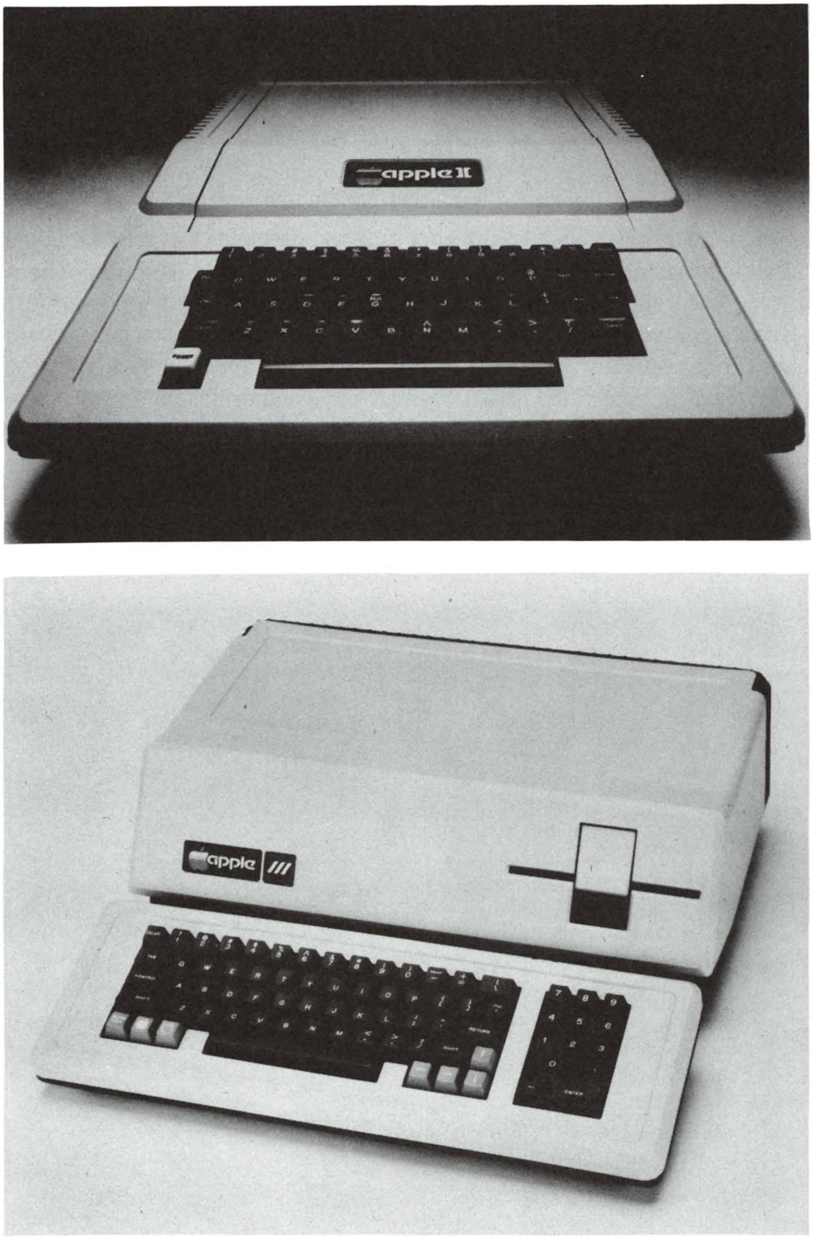
(a)

(b)

Figure 6-5. (a) Heath/Zenith H-8 system; (b) internal view. (Courtesy of Heath/Zenith)

Also built into the Apple II is a video display circuit that will drive an ordinary television monitor. The regular video format is 40 characters per line, with a total of 24 lines on the CRT screen at any one time. An interesting feature of the Apple II video monitor is that you can use either regular (white characters on black background) or inverse (black characters on white background) modes, and can cause some of the characters to flash on and off. The color graphics video display is capable of 15 different colors on a normal color video monitor.

A high-resolution video display provides 280h × 192v capability, allowing the programmer to draw graphs and other displays on the CRT screen.



(b)

Figure 6-6. (a) Apple II; (b) Apple III. (Courtesy of Apple Computers)

Heath H-11

The Heath H-11 microcomputer is based on the Digital Equipment Corporation LSI-11 16-bit microcomputer board. This machine is fully compatible with the famous DEC PDP-11 series of minicomputers and has a lot more capability than most microcomputers. When interfaced with a disk drive system (see Figure 6-7), the H-11 machine is compatible with the PDP-11/03, although at a much lower cost. In that configuration, the H-11 has 56K bytes of on-board RAM and up to 500K bytes of disk storage in IBM 3740 format.

Most users will not need a machine such as the H-11, but it is of immense use to those who have a large number of program pacakges that are based on the DEC PDP-11 machines (of which there are many in current use). It might be wiser for some users to go ahead and buy the H-11 instead of some lower-cost microcomputer that could also do a limited job, if it would mean that no software rewrite would be necessary. The costs of software in any computer system frequently, in fact usually, exceed the hardware cost; some authorities put the cost of software from commercial sources at $200 per line of working (i.e., debugged) program.

The Heath H-11 has available several different interfacing and accessory options, such as serial interface (serial RS-232 or 20-mA current loop), parallel

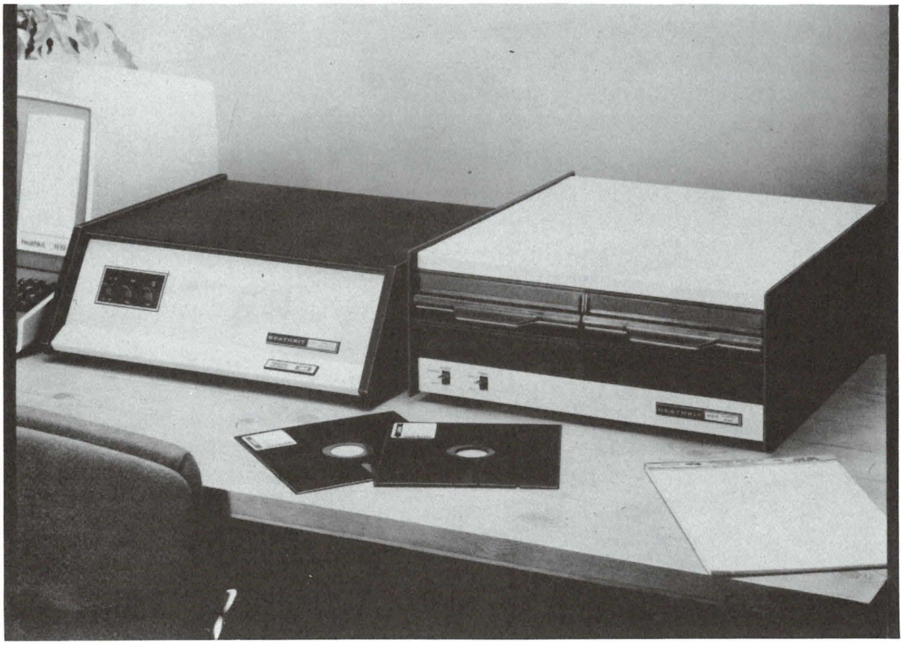


Figure 6-7. Heath H-11 system. (Courtesy of Heath/Zenith)

I/O, hardware multiplication/divide, and a wire-wrapping board for breadboarding circuits.

6-6 MICROPROCESSOR FUNDAMENTALS

The microprocessor chip revolutionalized the electronics industry. Although initially thought of as either a small logic controller or as a data-processing machine (depending upon your perspective and the first chip you saw), the microprocessor developed into a major force with hundreds or even thousands of applications in less than a decade.

What is a microprocessor? How does it relate to microcomputers? In this chapter we will explore these questions and give you a good grounding in computer technology basics. But first we will study computers in general by describing a typical programmable digital computer in block diagram form.

Mythical Analytical Device

Rather than mold our discussion around any one manufacturer's product, let's make up one that is sufficiently general to cover a large number of actual devices. Our computer will be nicknamed the *mythical analytical device*, or MAD, because the acronym often adequately describes both the emotional state of frustrated programmers and the mental health state of computer science buffs.

Figure 6-8 shows the block diagram of MAD. Let's first describe its parts and then describe a typical program operation.

MAD, like any programmable digital computer, has three main parts: *central processing unit* (CPU), *memory*, and *input/output* (I/O). There are certainly other functions in specific machines, but many of these are either special applications of these main groups or are too unique to be described in a general machine.

The central processing unit controls the operation of the entire computer. It consists of several necessary subsections, which will be described in greater detail later.

Memory can be viewed as an array of cubbyholes such as those used by postal workers (Figure 6-9) to sort mail. Each cubbyhole represents a specific *address* on the letter carrier's route. An address in the array can be uniquely specified (identifying only *one* location) by designating the *row* and *column* in which the cubbyhole is found. If we want to specify the memory location (i.e., cubbyhole) at row 3 and column 2, we could create a *row X column* address number, which, in this example, would be 32.

Each cubbyhole represents a unique location in which to store mail. In the computer, the memory location stores a binary word of information. In an 8-bit computer, each location will store a single 8-bit binary word.

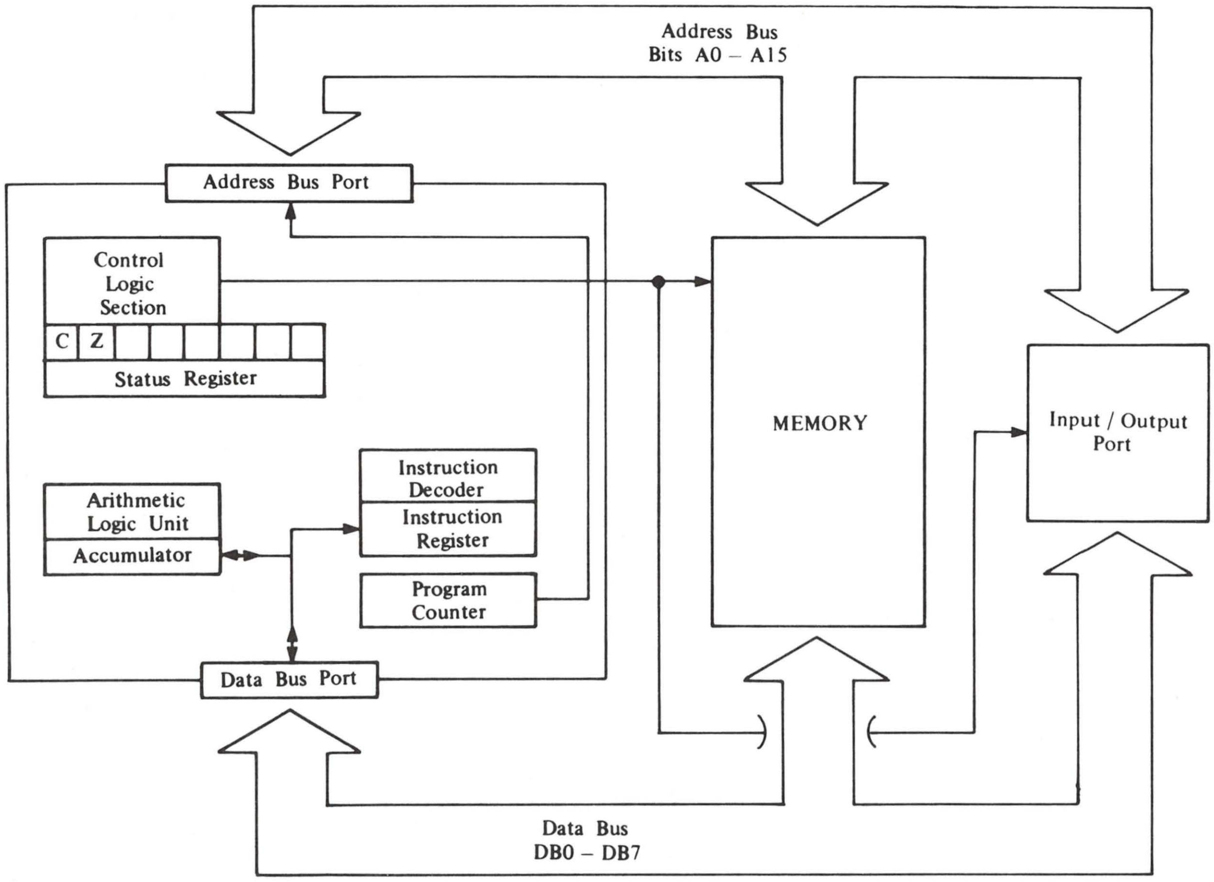


Figure 6-8. Block diagram of typical computer.

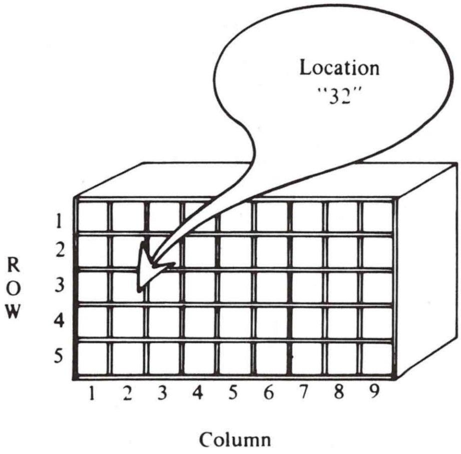


Figure 6-9. Analogy of computer memory.

There are three lines of communication between the memory and the CPU: *address bus, data bus* and *control logic signals*. These avenues of communication control the interaction between memory I/O, on the one hand, and the CPU on the other. Ultimately, therefore, they also control the functioning of the entire computer.

The *address bus* (bits A0 through A15 in Figure 6-8) communicates to the memory bank the address of the exact memory location being called by the CPU, regardless of whether a read or write operation is taking place. The address bus consists of parallel data lines, one for each bit of the binary word that is used to specify the address location. In most 8-bit microcomputers, for example, the address bus consists of 16 bits. A 16-bit address bus can uniquely specify 2^{16} or 65,536 different locations. This size is called "64K," not "65K" as one might expect; the reason is operative. Lowercase "k" represents the metric prefix *kilo*, which denotes 1000. Since 2^{10} is 1024, however, computer users decided that the kilo would be 1024 not 1000. But an uppercase "K" is used, rather than the "k" used for the metric kilo.

The size of memory that can be addressed *doubles* for every bit added to the length of the address bus. Hence, adding 1 bit to our 16-bit address bus creates a 17-bit address bus, which can designate up to 128K locations. Some 8-bit machines that have 16-bit address buses can be made to look like bigger machines by certain tactics that make a longer pseudo-address bus. In those machines, several 64K memory banks are used to simulate continuously addressable 128K, 256K, or 512K memories.

The *data bus* is the communications channel over which data travel between the main register (called the *accumulator* or *A-register)* in the CPU and the memory. The data bus also carries data to and from the various input and/or output ports. If the CPU wants to read the data stored in a particular memory location, those data are passed from the memory location over the data bus to the accumulator register in the CPU. Memory write operations are in exactly the opposite direction, but are otherwise the same.

The size of the data bus is usually cited as the size of the computer. An 8-bit microprocessor/microcomputer, therefore, is one that has an 8-bit data bus; a 16-bit microcomputer will have a 16-bit data bus. Do not be confused by a salesperson who might tell you that a 6502-based machine (8-bit data bus) is "in reality" a 16-bit machine because it has a 16-bit address bus.

The last memory signal is the *control logic* or timing signal. These are one or more binary logic signals that tell memory *if* it is being addressed and whether the request is a read or write operation. The details of control logic signals differ between different microprocessor chips, and so will not be discussed here. Typical signals are those of the 6502 and Z80 machines.

The *input/output* (I/O) section is the means by which the CPU communicates with the outside world. An input port will bring data in from the outside world and then pass them over the data bus to the CPU, where they are stored in the accumulator. An output port reverses that data flow direction.

In some machines, there are separate I/O instructions that are distinct from memory instructions. The Z80 is one such machine. The Z80 will pass the port address over the lower 8 bits of the 16-bit address bus (the 8-bit I/O address used in the Z80 can uniquely address up to 256 different ports). In other machines, such as the 6502, there are no distinct I/O instructions. In those machines, the I/O components are treated as memory locations; this technique is called *memory mapping* or *memory-mapped I/O*. Input and output operations then become memory-read and memory-write operations, respectively.

Central Processing Unit (CPU)

The CPU is the heart and brains of any programmable digital computer, including MAD. Although there are some different optional features in certain machines, all will have the features shown in our MAD computer (Figure 6-8). The principal subsections of the CPU include (at least) the following: *accumulator* or *A-register*, *arithmetic-logic unit* (ALU), *program counter* (PC), *instruction register*, *status register*, and *control logic section*.

The accumulator is the main register in the CPU and will have the same bit length as the data bus. All instructions executed by the CPU involve data in the accumulator, unless otherwise specified in the description of an instruction. Therefore, an ADD instruction causes an arithmetic addition of the data cited by the instruction to the *contents of the accumulator*.

Although there are often other registers in the CPU, the accumulator is the main register. The main purpose of the accumulator is the *temporary* storage of data being operated on by the instruction being executed. Note that data transfers to and from the accumulator are nondestructive. In other words, data "transfers" are not really transfers at all, but are, instead, *copying* operations. Suppose, for example, the hexadecimal number $8F_{16}$ is stored in the accumulator when an instruction is encountered requiring that the contents of the accumulator be stored at memory location $A008_{16}$. After the instruction is executed, we will find $8F_{16}$ both in memory location A008 *and* in the accumulator. If we have the opposite operation (i.e., transfer contents of accumulator to location $A008_{16}$), then we will see the same situation; after the transfer, the data will be in both locations. Since the accumulator contents change every time an instruction is executed, we will have to use such transfers to hold critical data someplace in memory.

The arithmetic-logic unit (ALU) contains the circuitry that performs the arithmetic operations of addition and (sometimes) subtraction, plus the logical operations of AND, OR, and XOR.

The program counter (PC) contains the address of the next instruction to be executed. The secret to the success of a programmable digital computer is its ability to *fetch and execute instructions sequentially*. Normally, the PC will increment appropriately (1, 2, 3, or 4) while executing each instruction: 1 for 1-byte instructions, 2 for 2-byte instructions, and so on. For example, the instruction LDA,n is a 6502 instruction mnemonic that loads the accumulator with the number *n*. In a program, we will find the code for LDA,n followed by *n*.

Location	*Code*	*Mnemonic*
0100		LDA,n
0101	*n*	*n*
0102	(next instruction)	

At the beginning of this operation, PC = 0100, but after execution it will be PC = 0102 because LDA,n is a 2-byte instruction.

There are other ways to modify the program counter. Executing any form of JUMP instruction, for example, modifies the contents of the PC to contain the address of the jumped-to location. Another way to modify the PC contents is to activate the *reset* line. The computer sees reset as a hard-wired JUMP to location 0000.

The *instruction register* (IR) is the temporary storage location for instructor codes stored in memory. When the instruction is fetched from memory by the CPU, it will reside in the instruction register until the next instruction is fetched.

The *instruction decoder* is a logic circuit that reads the instruction register contents and then carries out the intended operation.

The *control logic section* takes care of housekeeping chores within the CPU and issues or responds to control signals from the outside world. These signals are not universally defined but control such functions as memory requests, I/O requests, read/write signaling, interrupts, and so forth.

The *status register*, also sometimes called *status flags*, is used to indicate to the program and (sometimes) to the outside world the status of the CPU at any given instant. Each bit of the status register represents a different function. Different microprocessor chips use slightly different status register architectures, but all will have a *carry flag* (C) to indicate when an instruction execution caused a carry, and a *zero flag* (Z) to indicate when an arithmetic or logic instruction resulted in zero or nonzero in the accumulator (typically, $Z = 1$ when the result is zero).

We have now developed the CPU for our MAD computer. This discussion in general terms also describes a typical microprocessor chip; a microprocessor (as opposed to a single-chip computer) is essentially the CPU portion of a MAD.

Operation of MAD

A programmable digital computer such as MAD operates by sequentially fetching, decoding, and then executing instructions stored in memory. These instructions are stored in the form of *binary numbers*. In some early machines there were two memories, one each for program instructions and data. The modern method, however, uses the same memory for both data and instructions.

How, one might ask, does the computer know whether the binary number stored in any particular location is an instruction, data, or an alphanumeric character representation (e.g., ASCII or BAUDOT codes)? The answer to this important question is the key to the operation of MAD: *cycles*.

The MAD operates in cycles. A computer will have at least two cycles: *instruction fetch* and *execution*, and these are often broken into several subcycles. The details differ even though general methods are similar.

Instructions are stored in memory as binary numbers called *operation codes* (op-codes, for short). During the instruction fetch cycle, an op-code will be retrieved from the memory location specified by the program counter and stuffed into the instruction register. The CPU assumes that the programmer was smart enough to arrange things such that an op-code will be stored at that location when the PC increments to that address.

During the first cycle, an instruction is fetched and stored in the instruction register. During the second cycle, the instruction decoder will read the IR and then carry out the indicated operation. When these two cycles are completed, an instruction will have been fetched and executed, the program counter incremented to reflect the memory location that will contain the next instruction, and the CPU made ready for the next instruction. The CPU will then enter the next instruction fetch cycle and the process repeats itself. This process continues over and over again as long as the MAD is working. Each step is synchronized by a train of clock pulses so that events remain rational.

From our description you might be able to glean a truth concerning what a computer can or cannot do. The CPU can shift data around, perform logical operations (e.g., AND, OR, XOR), and add two N-bit numbers (sometimes subtract as well), all in accordance with a limited repertoire of binary word instructions. These chores are performed *sequentially* through a series of discrete steps. The secret to whether a problem is amenable to computer solution depends upon whether a plan of action (called an *algorithm)* can be written that will lead to a solution by a sequentially executed series of steps. Most practical instrumentation, control, or data-processing chores can be so solved, a factor that accounts for the meteoric rise of the microprocessor. The field of endeavor that studies sequential solutions to practical (and some not so practical) problems is called *numerical methods*.

chapter 7

ELECTRONIC MULTIMETERS

7-1 OBJECTIVES

1. To learn the principles behind the operation of electronic multimeters.
2. To learn the different types of electronic multimeter.
3. To learn how electronic multimeters differ from the VOM.
4. To learn to use and interpret readings from the electronic multimeter.
5. To learn the precautions necessary when one is using electronic multimeters.

7-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. How does an FETVM differ from (a) the VOM? (b) the VTVM?
2. Describe in your own words the operation of an FETVM. Make a drawing if it will aid your discussion.
3. Describe the operation of a *dual-slope integrator* digital multimeter. Use a drawing if necessary.
4. How does a *true-rms computer* measure the rms value of an ac waveform?
5. What is meant by a "3½-digit" multimeter?

7-3 VOM-vs-EMM

The volt-ohm-milliammeter, or VOM, is a rugged and accurate instrument, but suffers from certain disadvantages. The main problem is that it lacks both *sensitivity* and *input impedance*.

The sensitivity of a VOM is given in the units ohms/volt (Ω/V), and has been described earlier as the reciprocal of the meter movement's full-scale current. A 50-microampere movement used with a multiplier resistor to make a voltmeter, for example, has a sensitivity of $1/5 \times 10^{-5}$, or 20,000 ohms/volt. On low voltage ranges, then, the impedance of such an instrument tends to be low. In the case of the 20,000 ohms/volt meter with a 0 to 0.5-volt range, the input impedance is only (0.5 V)(20,000 Ω/V), or 10 kohms.

The electronic voltmeter (EVM), on the other hand, can have an input impedance as low as 10 megohms, or as high as 100 megohms, and the input impedance is *constant* over all ranges instead of being different on each range as in the VOM. The loading of the circuit under test tends to be less with EVM than with the VOM.

On the negative side, however, many EVMs cannot be used in the presence of strong electric or electromagnetic fields such as those produced by a radio transmitter, television flyback transformers, and so forth. The field will tend to bias the transistors or integrated circuits used in the EVM to a point where they will not operate properly. One popular EVM model will read a constant one-third scale whenever the instrument is used inside an AM broadcasting station! The VOM, on the other hand, is practically immune to such effects except on the ac scales.

7-4 THE BASIC ELECTRONIC MULTIMETER

An electronic multimeter (Fig. 7-1) might use any of several technologies, including vacuum tubes, transistors, integrated circuits, or digital circuitry. Some meters use an input amplifier to increase the amplitude of weak signals, while others use an *attenuator* to reduce the amplitude of signals that are above the meter's basic range.

The electronic multimeter might be ac mains powered, or it might be battery powered to allow portability. While the power supplies may differ in type, the one commonality among all electronic multimeters is that they use active devices, so they require a power supply.

The Simpson model 314 shown in Fig. 7-1 measures full-scale ac voltages down to 10 mV, and dc voltages down to 50 mV. The current ranges are from 10 μA full scale to 1 ampere full scale. The ohmmeter ranges are from $R \times 1$ to $R \times 1$ meg. A function switch provides selection for ac volts, dc volts, ac and dc current and two ohms ranges, high and low (about which more later.).

Electronic voltmeters that use an analog meter movement almost invariably use an electronic balance circuit such as that shown in Fig. 7-2 to drive the meter.

The different types of electronic voltmeter (EVM) go by a variety of names that tend to reflect the technology used in the balance circuit. The original EVMs used vacuum tubes, so they were called *vacuum tube voltmeters,* while solid-state models with JFET input stages were known as *transistor voltmeters* (TVM)

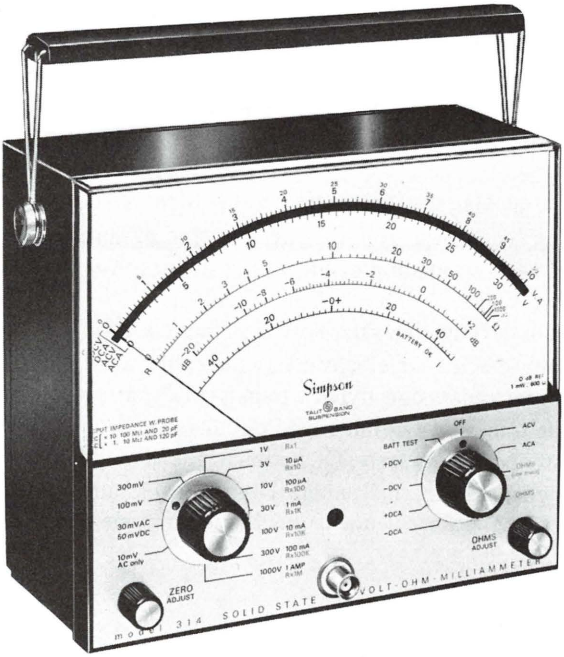


Figure 7-1. Analog Electronic voltmeter. (Courtesy of Simpson Electric.)

or *FET voltmeters* (FETVM). The basic circuit operates on the same principles, however, regardless of the technology employed.

The circuit in Fig. 7-2 uses a *differential amplifier* consisting of transistors $Q2$ and $Q3$ to form a balanced bridge circuit. Field effect transistor $Q1$ serves as a source follower and is used to provide impedance transformation between the input and the base of $Q2$.

Current $I1$ is generated by a *constant current source,* which may be a high value resistor to the V-source, or one of the electronic CCS sources that are known. Since $I1$ is constant, we know that the following relationship holds true.

$$I1 = I2 + I3 = k \tag{7-1}$$

The bias on $Q2$ and $Q3$ is such that $I2 = I3$ when E_x is zero. Under that condition $E_a = E_b$, so the current through dc meter movement $m1$ is zero; $I4 = 0$.

The bias on $Q3$ is fixed by a stable reference supply, but the bias on $Q2$ is a function of the voltage drop across resistor $R2$, which in turn, is controlled by E_x. When an unknown voltage E_x is applied, the bias on $Q2$ increases and that causes voltage E_a also to increase. Now, E_a is greater than E_b, so current $I4$ is no longer

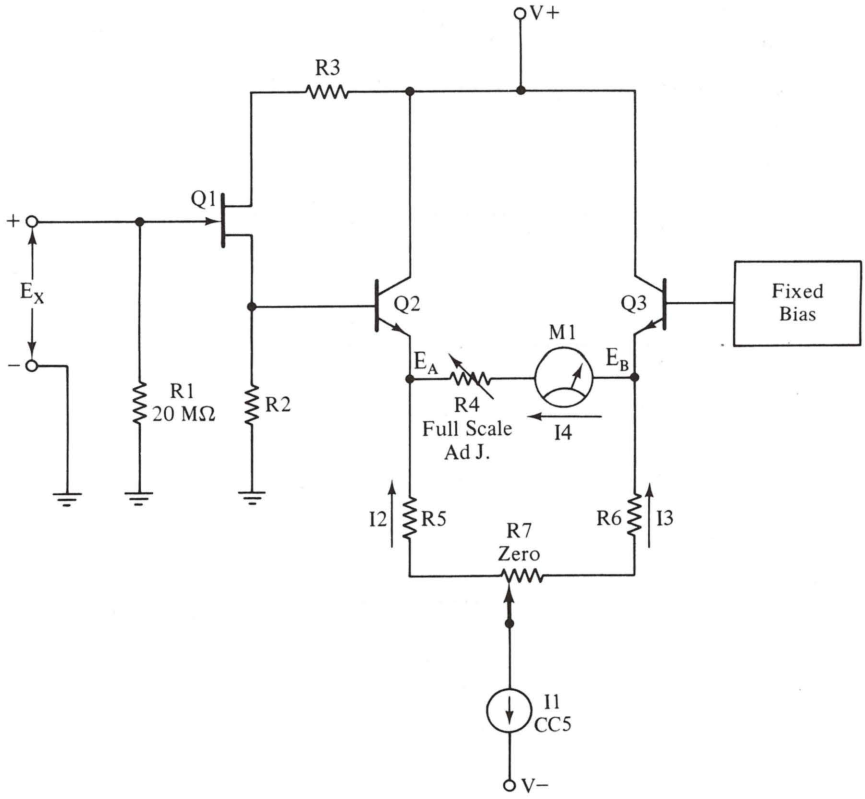


Figure 7-2. Typical solid-state electronic voltmeter circuit.

zero. The magnitude of current *I*4 flowing in *m*1, hence the deflection of the *m*1 pointer, is proportional to E_X.

The value of E_X that causes maximum deflection of *m*1 is the *basic range* of the instrument, and is usually the lowest range on the selector switch in nonamplified models. Higher ranges can be accommodated through the use of an *input attentuator* such as that shown in Fig. 7-3, while lower ranges can be accommodated by a preamplifier.

The attenuator in Fig. 7-3 is a resistance voltage divider. The full-scale voltage from the probe appears across the divider, so the voltage at each tap is a progressively lower fraction of the full input voltage. If the basic range of the balance circuit is 50 mV, then a 1000-volt input potential will be reduced to 50 mV when the selector switch is connected to the bottom tap. Precision resistors are used for *R*1 through *R*10, and most of these are of the *wirewound* variety. Such resistors have significant *inductance,* so if the meter is to be used on ac

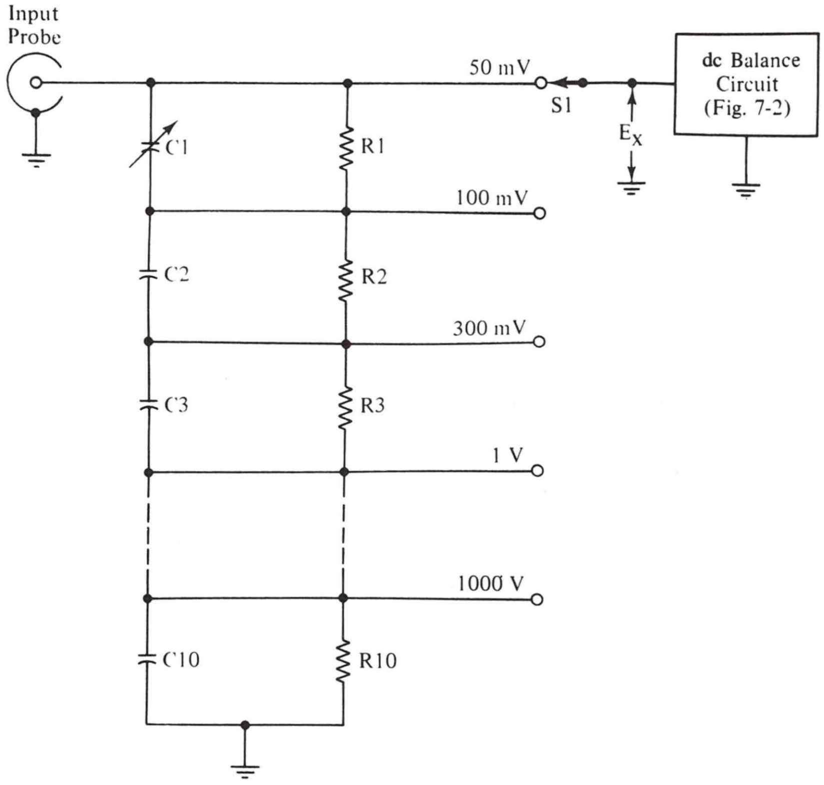


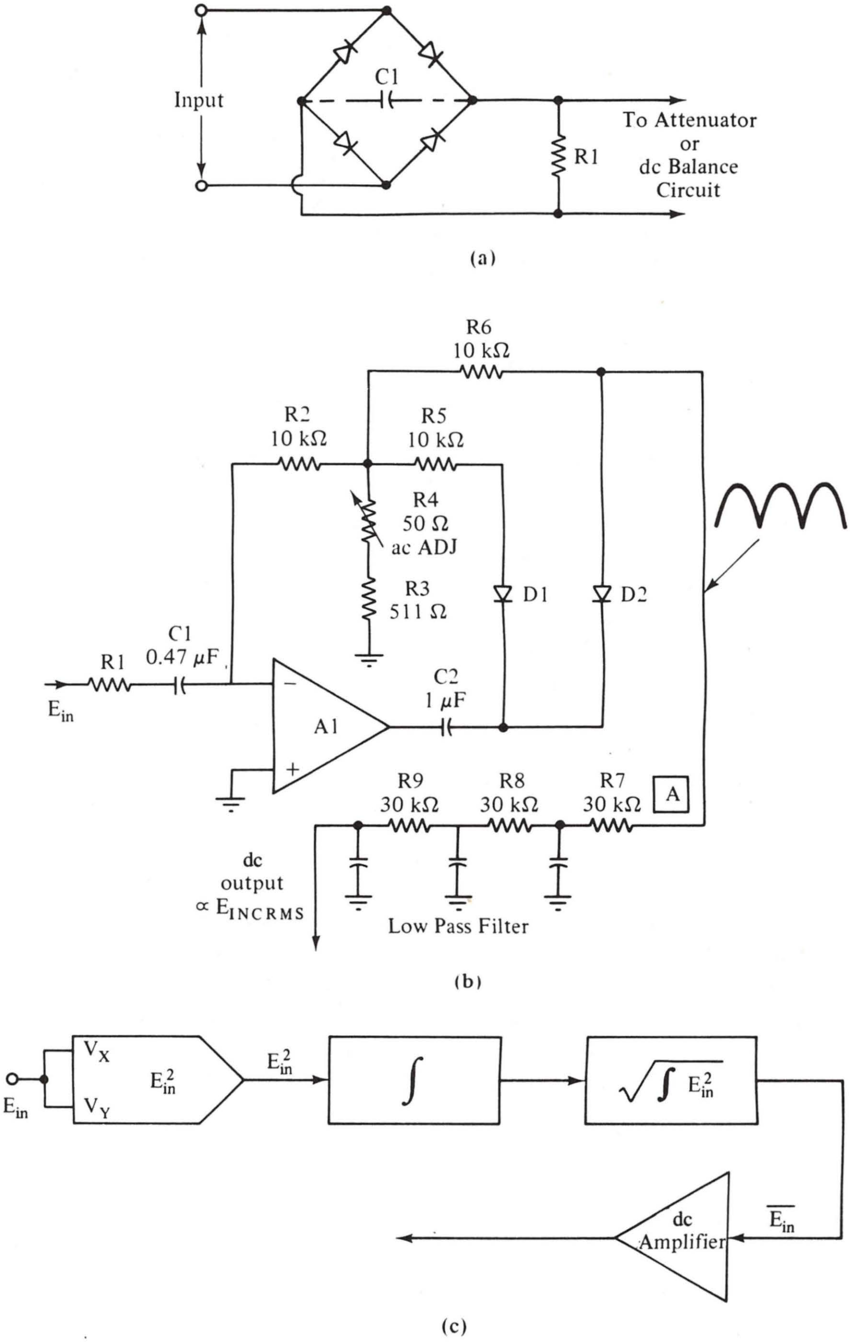
Figure 7-3. Input attenuator circuit.

some sort of compensation is required. The compensation takes the form of capacitors $C1$ through $C10$. Several of these capacitors are made variable, and these capacitors are adjusted, with a square wave applied to the input, for best "squareness" as viewed on an oscilloscope.

7-5 AC MULTIMETERS

Most basic EVM circuits such as that in Fig. 7-2 are for use with dc potentials only. To accommodate ac potential some sort of converter is required.

In the simplest case, shown in Fig. 7-4(a), a full-wave rectifier is used to drive the meter circuit. This type of converter tends to suffer from most of the same faults as passive ac voltmeters that use the same principle. Some circuits make the circuit essentially *peak reading* by including capacitor $C1$. This



(a)

(b)

(c)

Figure 7-4. (a) Simple ac rectifier. (b) Rms averaging circuit. (c) Computing-type rms converter.

capacitor will charge up to the peak value of the ac waveform due to the action of the rectifier.

A somewhat superior rectifier/averager circuit is shown in Fig. 7-4(b). In this case the rectification is provided by an operational amplifier circuit called an *ideal rectifier* or *precision rectifier*. A normal pn junction diode is nonlinear at low forward voltages. Silicon diodes do not become linear until the applied forward voltage exceeds approximately 0.6 to 0.7 volt, while germanium diodes require 0.2 to 0.3 volt of forward bias. In the circuit of Fig. 7-4(b) these nonlinearities are "servoed out" by the operational amplifier and its negative feedback loop.

The pulsating dc at point A in the circuit is smoothed out, i.e. *averaged,* by the low-pass filter $R7$–$R9/C3$–$C5$. The relatively pure dc output from the filter is applied to an analog balance circuit such as Fig. 7-2, or a dc digital voltmeter.

Although most ac voltmeters are calibrated in terms of *volts rms,* most will not actually read the rms value unless the input signal has a pure sine wave shape. The reasons for this were discussed in Chapter 2.

Some electronic voltmeters, however, use an *rms-to-dc converter* [block diagrammed in Fig. 7-4(c)] that computes the rms value of most common ac waveforms. Most such circuits operate to 10 or 20 kHz, while some operate to over 1 mHz. Recall the meaning of the term *rms* from Chapter 2:

$$E_{\text{rms}} = \sqrt{\frac{1}{T}\int_{0}^{T} (E_{\text{in}})^2 \, dt} \tag{7-2}$$

The circuit in Fig. 7-4(c) is an analog computer that solves this equation. The first stage is an analog multiplier connected as a *squarer* circuit; i.e., the V_x and V_y terminals are connected together, making the output proportional to $(E_{\text{in}})^2$. The second stage integrates the output of the squarer, and the third stage takes the square root of the integrator output. The output of the square root stage is proportional to the *true rms* value of the input waveform.

7-6 ELECTRONIC OHMMETERS

Ohm's law tells us that resistance in ohms can be computed from the current in amperes and the voltage, namely:

$$R = E/I \tag{7-3}$$

If the current through an unknown resistor is held *constant,* then the value of the voltage drop across the resistance will give us the data needed to calculate the resistance value. If that current is a power of ten submultiple of 1 ampere, i.e., 1 mA, 100 μA, etc., then the voltage drop will be numerically equal to the resistance. Only repositioning of the decimal point is then needed to read out directly in ohms or kilohms. Figure 7-5 shows such an ohmmeter circuit.

The constant current source (CCS) holds the current through R_X at a constant 1 mA (i.e., 0.001 ampere). By Ohm's law, then,

$R(\Omega)$	E (volts)
100	0.1
1000	1.0
10,000	10.0

The gain of the dc amplifier and the output current of the CCS are set by the range selector switch so that full-scale resistances from milliohms to megohms can be accommodated.

The low current levels used in this type of ohmmeter circuit confer both advantages and disadvantages in practical situations. Ordinary ohmmeters using passive dc components and meter movement use a 1.5-volt dc battery (or even larger), and these produce a relatively high current in low resistances. That type

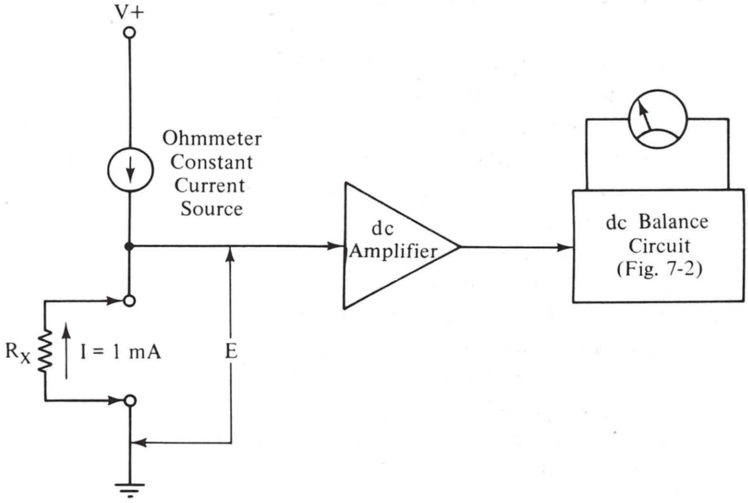
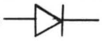


Figure 7-5. Electronic ohmmeter circuit.

of ohmmeter will forward-bias pn junctions, so it will result in large errors when used in solid-state circuits. But the low voltage levels used in the circuit of Fig. 7-5 will not usually forward-bias pn junctions, so it may be used in solid-state circuits.

The inability to forward-bias pn junctions is also a disadvantage, because forward and reverse resistance readings across the junctions often serve as a quick test for diodes and transistors. Some EVM manufacturers include a high-power ohmmeter scale so that such tests can be made. In meters such as the Simpson 314 shown in Fig. 7-1, high- and low-power ohmmeters are provided at different settings of the function switch, whereas in other products a pushbutton switch is provided marked "diode" or with the  symbol.

7-7 DIGITAL VOLTMETERS

A digital voltmeter uses an analog-to-digital converter (ADC) to convert the dc input, or output of an ac converter circuit, to a *binary coded decimal* (BCD) digital word that is used to drive a digital display device. Most digital voltmeters (DVM) or digital multimeters (DMM) use an ADC circuit called the *dual-slope integrator,* which we will now discuss in detail. Other ADC circuits are used in certain other instrumentation applications, and these are further discussed in Chapter 14.

7-8 DUAL-SLOPE INTEGRATION

The block diagram of a dual-slope integrator is shown in Fig. 7-6(a), while associated waveforms are shown in Fig. 7-6(b).

The heart of the circuit is the operational amplifier *integrator* (see Chapter 12) consisting of operational amplifier $A1$, plus $R1$ and $C1$, plus a *voltage comparator* ($A2$). The output of the comparator will remain LOW if the integrator output is zero, and HIGH if the integrator output is more than a few millivolts above ground potential.

At the beginning of the conversion cycle, the control logic section momentarily closes electronic switch $S2$ so that the charge on capacitor $C1$ goes to zero, and also ensures that $S1$ is set to position "A."

If $S1$ is in position "A," the integrator input is connected to the input voltage source, causing the voltage at the integrator output to begin rising [see time t_0 in Fig. 7-6(b)]. As soon as E_A rises a few millivolts the comparator output snaps HIGH, enabling the gate to pass clock pulses to the digital counter (Chapter 6) section. The counter is allowed to overflow [$t1$ in Fig. 7-6(b)], and the output *carry* pulse from the counter is used to tell the control logic section to switch $S1$ to position "B." This action connects the integrator input to a *precision refer-*

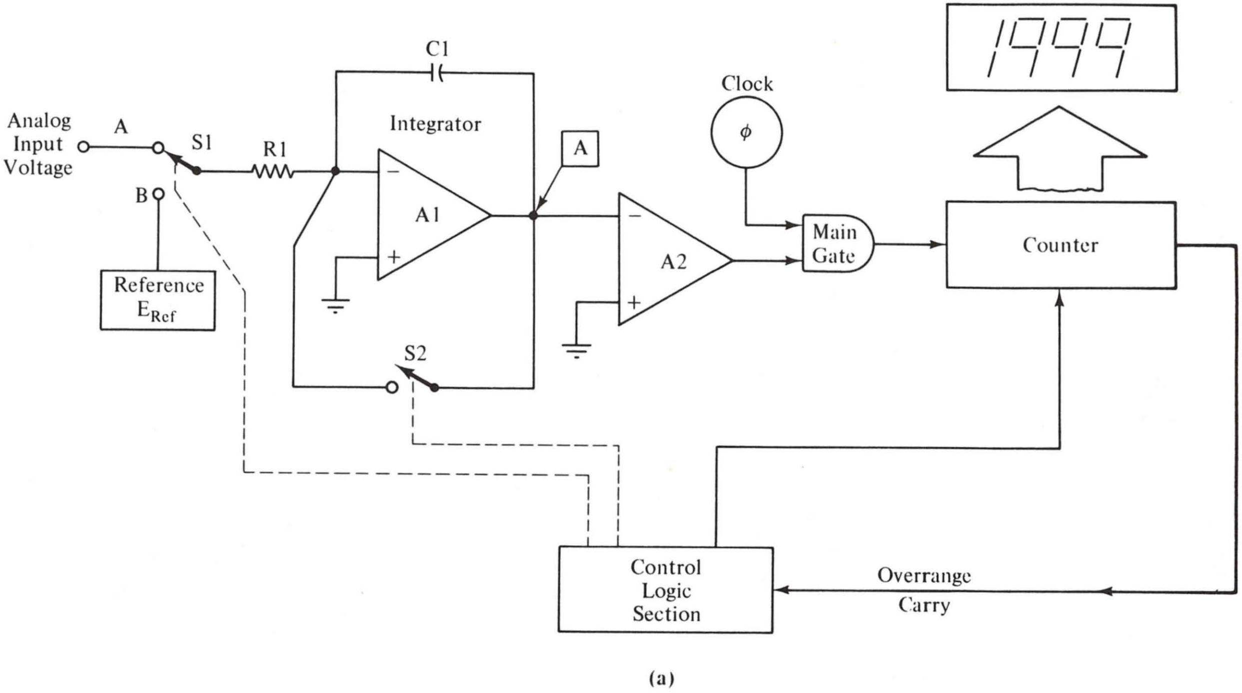


Figure 7-6. (a) Dual-slope integrator for electronic voltmeters.

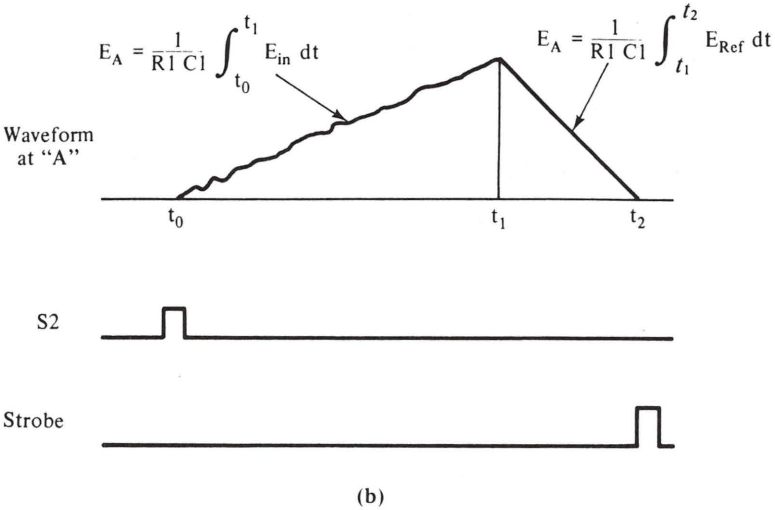

$E_A = \frac{1}{R1\,C1}\int_{t_0}^{t_1} E_{in}\,dt$
$E_A = \frac{1}{R1\,C1}\int_{t_1}^{t_2} E_{Ref}\,dt$
Waveform at "A"
t_0
t_1
t_2
S2
Strobe
(b)

Figure 7-6. *(cont.)* (b) Timing of the dual-slope integrator.

ence voltage source. The polarity of the input current created by the reference is such that it begins to *discharge* the integrator capacitor at a *constant* rate. The counter, meanwhile, has continued to increment, passing through 0000 at time $t1$, and continuing to accumulate clock pulses until E_A is back down to zero.

The value of E_A at time t_1 was proportional to the value of E_{in}. At the same instant the count was 0000. Since the counter continues to increment as the integrator discharges (t_1 to t_2), the *count* at the instant the gate is closed (when $E_A = 0$) is also proportional to E_{in}. By correct scaling the count will be numerically the same as the potential applied to the input. To recapitulate:

1. At time t_0 switch $S2$ is closed briefly to dump any residual charge in $C1$, and $S1$ is set to position "A."
2. The integrator begins to charge due to current $E_{in}/R1$, so E_A begins rising from zero.
3. As soon as E_A is greater than zero, the output of the comparator goes HIGH, enabling the main gate to pass clock pulses into the counter.
4. The counter increments until it overflows at time t_1, and the overflow pulse causes $S1$ to switch to position "B," applying the reference voltage to the input of the integrator. At this instant the count is 0000.
5. From times t_1 to t_2 the integrator *discharges* under the influence of current $-E_{ref}/R1$; meanwhile, the counter continues to increment.
6. At time t_2 the comparator shuts off the flow of clock pulses through the gate. The count accumulated between t_1 and t_2 represents the input voltage.

7. The control logic section issues a *strobe* pulse at time t_2 to update the counter's digital display.

The dual-slope integrator is one of the slower types of ADC circuit, but offers several advantages when used in DVM circuits: relative immunity to noise riding on the input voltage, relative immunity to error due to inaccuracy or long term drift in the clock frequency, etc. The 10 to 40 ms conversion cycle time poses no disadvantage in most DVM applications.

7-9 WHAT DOES THE DUAL-SLOPE DVM MEASURE?

A dc digital voltmeter tends to read the *average* input voltage over the period of integration. This feature makes the DVM particularly useful for measuring noisy signals. It also tends to average the pulsations from a rectifier-type ac-to-dc converter, provided that $t_2 - t_0$ is greater than about 18 ms.

The accuracy of a digital voltmeter, or DMM based on a DVM, can be quite good. In fact, it can exceed the accuracy of analog multimeters. You will see terms such as "2½-digit," "3½-digit," and "4½-digit" in the advertisements for DVM/DMM products, and will wonder what is meant by a "½" digit. This term refers to the fact that the *most significant* digit can only be a "0" or a "1," while all other digits can be anything between "0" and "9." Such terminology indicates that the meter can read 100 percent overrange from its basic range. For example, a 3½-digit DVM will read 0–1999 mV, while its basic range is only 0–999 mV. If this range is exceeded, then the "1" lights up; otherwise it remains darkened.

7-10 STEPPER-TYPE DVMS

Some of the original DVM instruments used a *stepper* technique instead of an integrator, and certain modern instruments have adapted the technique to electronic methods. In both cases, a voltage comparator is used to compare the input voltage with a *staircase* (i.e., quasi-ramp) reference voltage.

In the older type of instrument the staircase voltage was generated by a pulsed stepper relay that selected successively higher taps on a resistor voltage divider. The same pulse that moved the stepper relay also increments a digital events counter, so if each step represents an equal voltage change, then the counter reading will indicate the reference voltage value. For example, each step could represent 10 mV, so an input voltage of 1 volt would be reached when the stepper had incremented 100 steps (100 steps × 10 mV/ step is 1 volt).

Modern versions of this circuit replace the slow, noisy, stepper relay with a

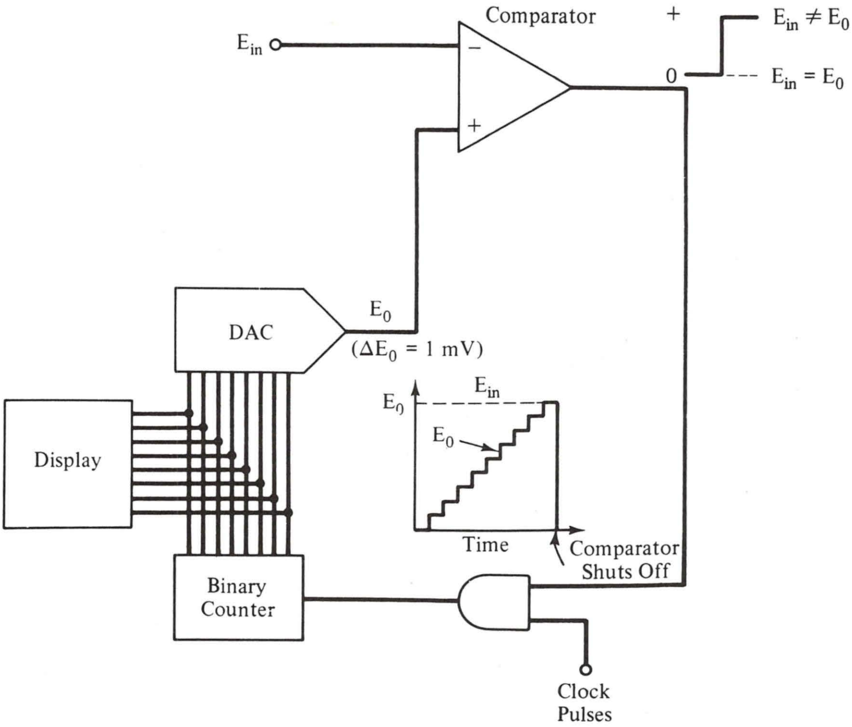


Figure 7-7. Stepper-type digital voltmeter circuit.

precision *digital-to-analog converter* (DAC) (Fig. 7-7) that does essentially the same job (see Chapter 14). The operation of the circuit is as follows:

1. At the beginning of the conversion cycle $E_0 = 0$, and E_{in} is some potential between 0 and 1 volt dc.
2. Since $E_0 \neq E_{in}$, the comparator output is HIGH, and this enables the gate to pass clock pulses to the counter.
3. DAC output voltage E_0 increases 1 mV per step until $E_0 = E_{in}$.
4. When $E_0 = E_{in}$, the comparator output drops LOW, turning off the gate. The output of the counter is now numerically equal to E_{in} in millivolts, and this value is displayed on the digital output indicator.

7-11 COMMERCIAL DIGITAL MULTIMETERS

The digital multimeter was, at one time, too expensive for most users. They were found in certain military, scientific, and commercial laboratories. Only a few companies, notably Fluke and Hewlett-Packard, produced the instruments. But today, integrated circuit technology has allowed the production of DVM and DMM instruments at previously unheard-of prices. One model based on a popular LSI chip sells for less than $50 in kit form, and the prices go up from there to several thousands of dollars. Do not expect a lot of accuracy or special features on a low-cost model, although very good performance is available in the medium-price range.

Figure 7-8 shows DMM products representative of a larger number of instruments. The model shown in Fig. 7-8(a) is the Hewlett-Packard model 3476A, and is a portable instrument designed for service technicians, field engineers, and so forth. It is battery powered, although it is able to operate from ac mains power if the charger cord is plugged in.

The H-P 3476A will measure ac volts 3.3 mV to 700 V_{rms}, dc volts from 100 μV to 1000 V, dc current from 100 μA to 1.1 A, ac current from 3.3 mA to 1.1 A, and resistance from 1 ohm to 11 megohms.

This particular model features *autoranging* operation, which means that the circuit will automatically adjust to the correct voltage, current, or resistance range. Other models by other manufacturers, but in the same price class, offer pushbutton selection of the ranges.

A pushbutton portable DMM is shown in Fig. 7-8(b). This type of instrument is based on an LSI chip, and is intended for field service applications.

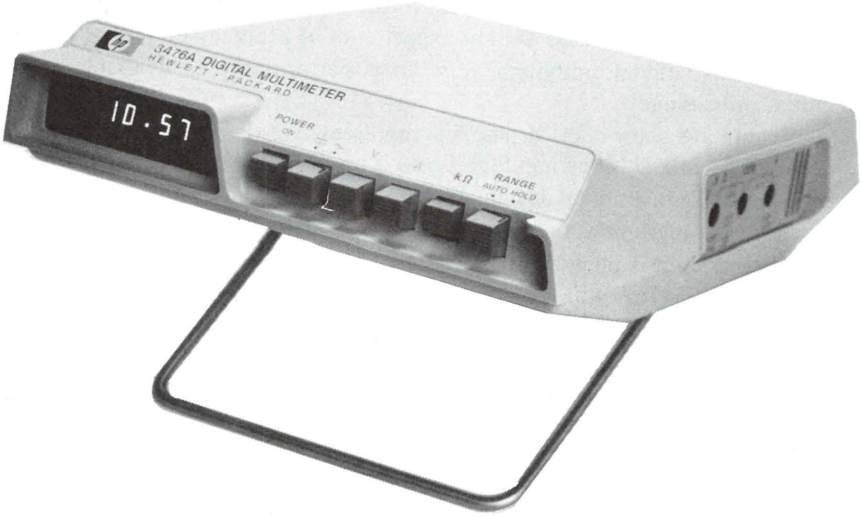
The Tektronix model DM-502 plug-in DMM is shown in Fig. 7-8(c). This instrument is part of Tektronix' versatile line of main-frame plug-in instruments that are designed to be installed in a portable or bench model instrumentation main frame/power supply. This line allows the user to equip a laboratory bench or field service engineer with the needed instruments in a conveniently packaged form.

At the high end of the price range are precision instruments such as the Hewlett-Packard Model 3455A shown in Fig. 7-8(d). This instrument is a very accurate model used mostly in precision laboratories. The 3455A is able to perform certain product and ratio measurements in addition to straight voltage determinations.

A new category of DMM is the programmable Tektronix Model DM-5010 shown in Fig. 7-8(e). This instrument is not just a high performance DMM, but is also capable of working in a general purpose interface bus (GPIB) IEEE-488 system.

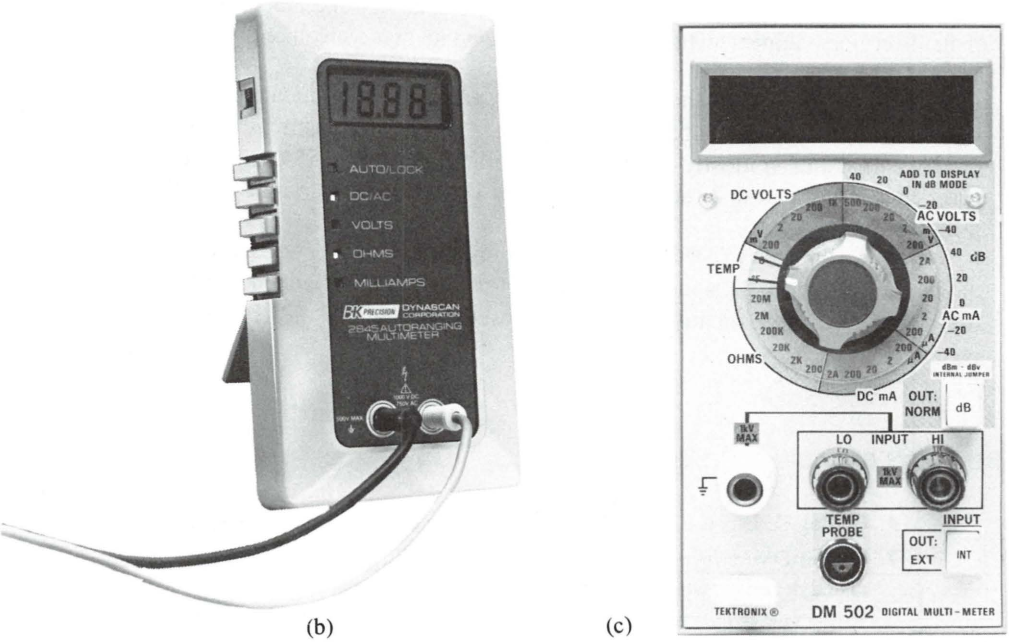
7-12 SUMMARY

1. Analog DMMs use a dc balance circuit to drive the meter movement.
2. Digital DMMs use an analog-to-digital converter to produce a binary coded decimal "word" that is proportional to the applied input voltage.
3. Resistance is measured by passing a current through the unknown resistance, and then measuring the voltage drop across the resistor.
4. High voltage ranges are accommodated through the use of an attenuator,



(a)

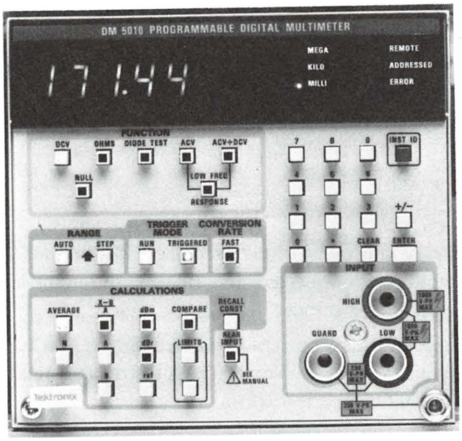
Figure 7-8. (a) Portable digital multimeter. (Courtesy of Hewlett-Packard.)



(b) (c)

Figure 7-8. *(cont.)* (b) Hand-held autoranging digital multimeter. (Courtesy of B&K Dynascan Corporation) (c) Mainframe electronic digital multimeter. (Courtesy of Tektronix, Inc.)

(d)



(e)

Figure 7.8. *(cont.)* (d) Laboratory-grade electronic digital multimeter. (Courtesy of Hewlett-Packard) (e) Tektronix DM-5010 digital multimeter. (Courtesy of Tektronix, Inc.)

while low voltage ranges are accommodated through the use of an amplifier.

5. EVMs can be more accurate than a VOM, and are more sensitive. EVMs tend to have a very high, and constant, input impedance.

7-13 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread the appropriate sections.

7-14 QUESTIONS

1. How does an analog EVM differ from a VOM?
2. Draw a simple dc meter balance circuit and describe its operation in your own words.
3. Name two types of analog-to-digital converters used in DVM designs.
4. What technique is used to measure resistance in many EVMs?
5. What is the difference between high- and low-power resistance circuits? Give at least one advantage to each type.
6. What is a "3½-digit" voltmeter? What is meant by a "½-digit"?
7. What is the resolution of a voltmeter described as a "4½-digit" instrument if the basic range is 0–9999 mV?
8. Can a 2½-digit DVM measure 1 m V if the basic range is 0–99 volts?

7-15 PROBLEMS

1. A 100-μA constant current is passed through an unknown resistance, and a voltmeter connected in parallel with the meter measures 888 mV. Calculate the resistance.
2. A 10-μA constant current is passed through a 100-kΩ resistor. What would a voltmeter connected across the resistor read? Assume an infinite input impedance on the voltmeter.
3. A DVM measures direct current by passing the current through a low-value fixed resistor, and then measuring the voltage drop across the resistor with a dc voltmeter. Find the current if the voltage drop across a 1-ohm resistor is 200 mV.

chapter 8

THE OSCILLOSCOPE

8-1 OBJECTIVES

1. To learn the principles of operation behind the cathode ray oscilloscope.
2. To learn the different types of oscilloscope.
3. To learn how to use and operate the oscilloscope.
4. To learn how to interpret oscilloscope waveforms.

8-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Name two types of CRT deflection system.
2. Draw a block diagram to a simple free-running sweep CRO.
3. Define in your own words (a) triggered sweep and (b) delayed sweep.
4. What are Lissajous figures? How are they produced?

8-3 THE CATHODE RAY OSCILLOSCOPE

The cathode ray oscilloscope (also called CRO, oscillograph, oscilloscope, or simply " 'scope") has been called the single most important instrument in electronics. It can be used to measure voltage (ac or dc), frequency, and to view the *waveshape* of the input signal.

The CRO is much like a television set in that it produces a picture of the shape of an input signal on a light emitting phosphor screen. The heart of the CRO is a special vacuum tube called the *cathode ray tube* (CRT).

8-4 CATHODE RAY TUBES

The operation of an oscilloscope depends upon the CRT, an example of which is shown in Fig. 8-1(a). The principal components of the CRT are the fluorescent phosphor screen, the vertical and horizontal deflection plates, accelerating electrode, focus electrode, and an electron gun. Some CRT models also have a post deflection accelerating electrode (sometimes called a *second anode*).

The electron gun is a specially constructed heated cathode that gives off electrons by thermionic emission. The cathode is surrounded by a negatively charged cylindrical control electrode that has a hole in one end. Electrons also carry a negative charge, so are repelled from the walls of the cylinder, and will stream out through the hole toward a less negatively charged *accelerator electrode*. In some CRTs this electrode is called anode 2, but beware of similar terminology when the post deflection accelerator is introduced.

The effect of the accelerator anode is to increase the velocity of the electrons, thereby increasing their kinetic energy. When the electrons strike the phosphor screen, they give up their kinetic energy in the form of *light energy*.

The beam width emitted from the electron gun is too broad for practical applications, so some focusing is required. This job is performed by the first anode, also called the *focus electrode*. A more detailed illustration of the focus system is given in Fig. 8-1(b). Note that both first and second anodes work together to form, in effect, an electron lens not dissimilar in function to an optical lens. Electric field lines exist between points on each of these electrodes. The electrons of the beam are affected by this field in such a way that they are forced to converge on a distant point external to the focus assembly. The distance between the convergence point and the second anode is determined primarily by the potential existing between the two anodes. It is usually the practice of oscilloscope designers to hold the high voltage applied to the second anode constant, and then to use a potentiometer labeled *focus* to vary the potential applied to the first anode. In practice, the operator adjusts the focus control for the sharpest dot on the CRT viewing screen, as shown in Figs. 8-1(c) and 8-1(d).

The basic CRT power supply circuit is shown in Fig. 8-2. Note that this example is but one of several alternative designs. Other models will probably use a modified version of this configuration.

A −1500-volt dc power supply is connected to a resistor voltage divider network consisting of $R1$ through $R3$. Potentiometer $R3$ varies the potential applied to the control electrode, designated as grid 1. This control serves the function of *intensity control*.

Potentiometer $R2$ serves as a *focus control,* and it varies the potential applied to the focus electrode, designated as anode 1. The potential applied to the accelerator electrode is kept constant.

It is interesting to note that the names given to the structures inside a CRT are referenced to structures of an ordinary vacuum tube; i.e., *anode, grid,* etc. In

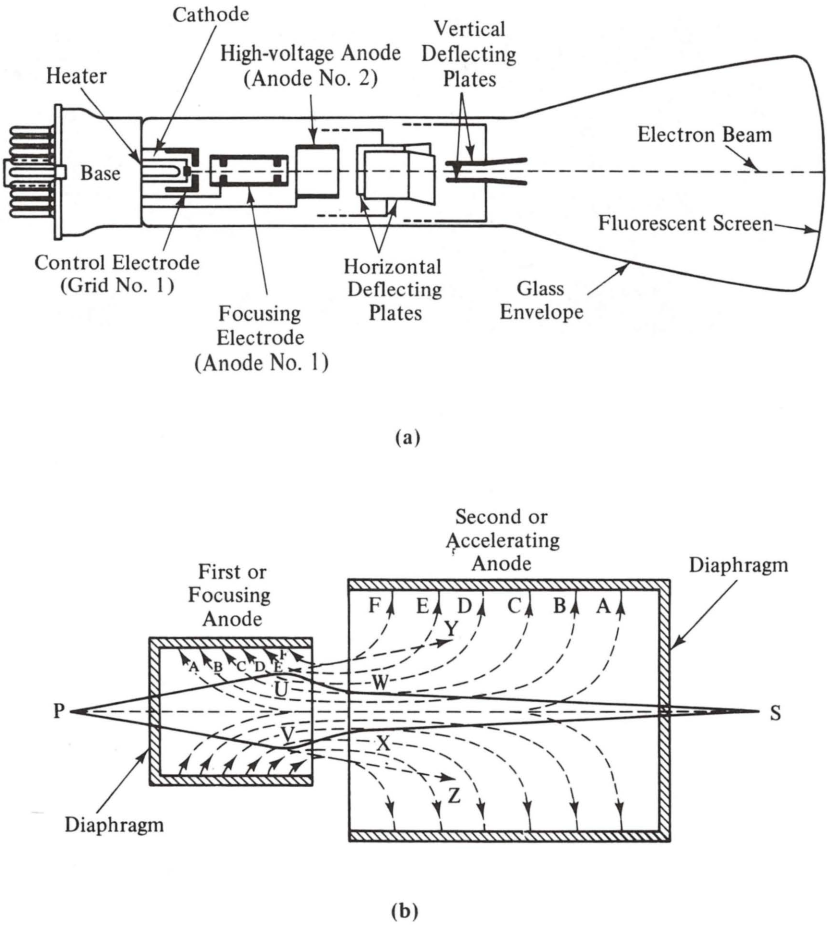
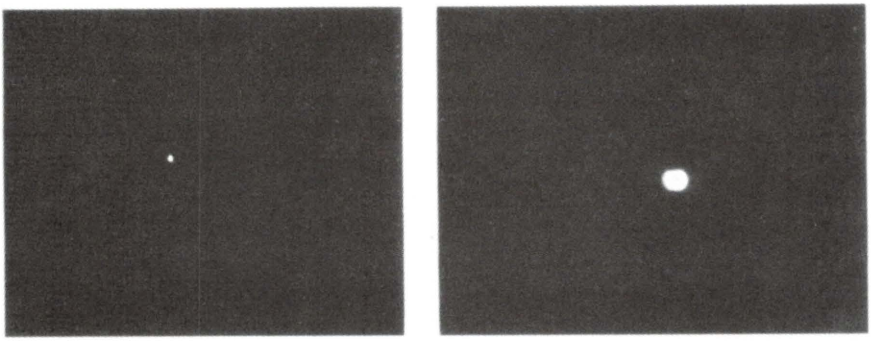

Cathode
Vertical Deflecting Plates
High-voltage Anode (Anode No. 2)
Heater
Electron Beam
Base
Fluorescent Screen
Control Electrode (Grid No. 1)
Horizontal Deflecting Plates
Glass Envelope
Focusing Electrode (Anode No. 1)
(a)
Second or Accelerating Anode
First or Focusing Anode
Diaphragm
F
E
D
C
B
A
Y
W
U
V
X
Z
P
S
Diaphragm
(b)

Figure 8-1. (a) Construction of the cathode ray tube. (b) Focus electrodes.

Fig. 8-1(a) we label certain structures as anodes, yet the schematic representation in Fig. 8-2 clearly shows them as *grids*. This seeming contradiction may tend to confuse you, unless you break the habit of using the vacuum tube as your frame of reference. The use of grid terminology here is merely a graphic device to simplify the circuit diagrams. If the grid symbols were not used, we would be required either to draw a picture of the symbols, which would be clumsy, or to invent a totally new symbology.

It is in the treatment of the second anode, or accelerator, potential that many oscilloscope models differ. In the example of Fig. 8-2 the second anode is at a potential of zero volts; i.e., it is *grounded*. Of course, being grounded, it is at a potential that is *more positive* than the other electrodes. In other models, the



(c) (d)

Figure 8-1. ***(cont.)*** (c) In-focus beam. (d) Out-of-focus beam.

second anode potential may be slightly positive, i.e., 0 to +150 volts, while in others it may be at a positive potential of several thousand volts. These situations are shown in Figs. 8-3(a) and 8-3(b), respectively. In the case of Fig. 8-3(a), the circuit is essentially the same as Fig. 8-2, except that a slight positive bias is applied to grid 3.

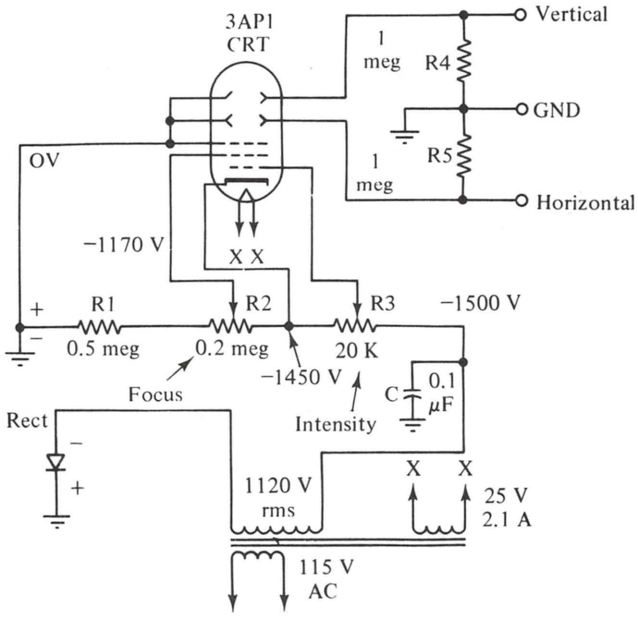
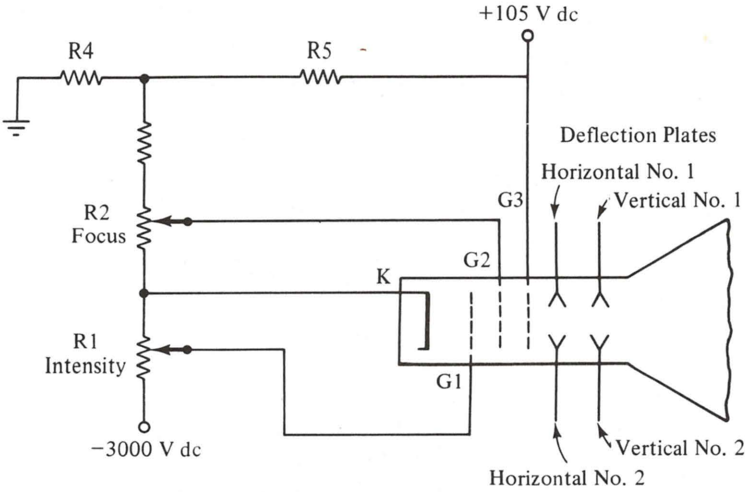
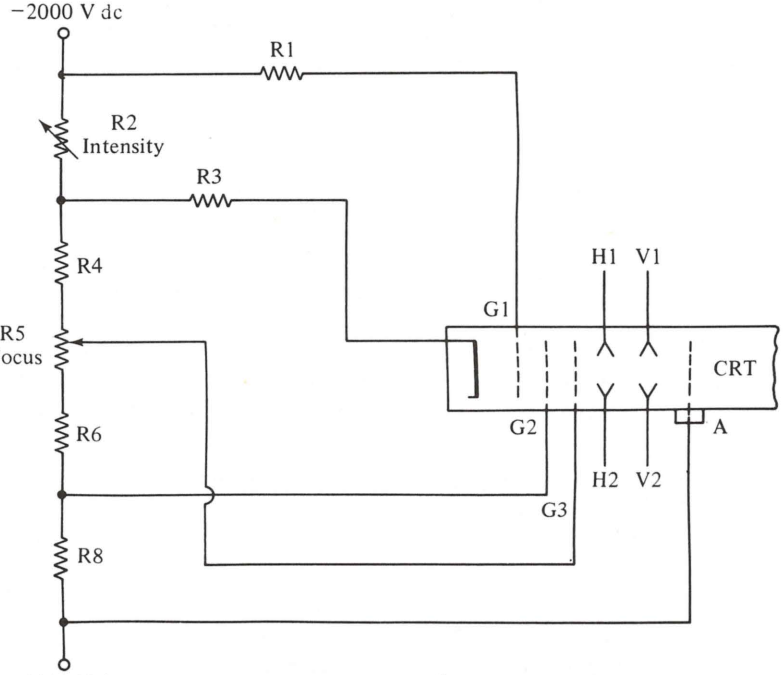


Figure 8-2. DC power supply circuit for CRT.



(a)



(b)

Figure 8-3. (a) Situation with second anode slightly positive. (b) Use of the post accelerator electrode.

In Fig. 8-3(b), on the other hand, another electrode is used past the deflection plates, closer to the phosphor screen, and is called the *post accelerator electrode*. The use of a post accelerator improves the linearity of the CRT. That is to say, the spot remains the same in a greater area of the viewing screen.

The post accelerator electrode is usually at a potential greater than 1 kilovolt in low-cost oscilloscopes, 3 to 7 kilovolts in higher priced 'scopes, and up to 25 kilovolts in color television receivers. Note that most CRTs using a post accelerator electrode bring the connection out of the glass envelope through a special high-voltage nipple, rather than through the socket with the other connections.

When the accelerated electron beam strikes the phosphorous screen, the electron's kinetic energy is given up. This energy is first absorbed by the phosphor atoms, but is then released as the electrons of those atoms return to ground state. This action causes the phosphor atoms to release light energy.

Some phosphors give up all absorbed energy rapidly, so the light emitted will diminish rapidly. Other CRTs use a different phosphor in which the emission of light is prolonged. The measure of the length of time a trace remains on the screen before fading is the *persistence* of the phosphor. Relatively short persistence phosphors are used in laboratory oscilloscopes, while longer persistence types are used in medical electronics, where the signals occur at low frequency rates.

8-5 DEFLECTION SYSTEMS

The *deflection system* of an oscilloscope moves the electron beam right and left, or up and down. There are two basic types of deflection system, *magnetic* and *electrostatic*.

In the magnetic system, there are no internal deflection plates, but an external electromagnet surrounds the glass neck of the CRT. There are actually two separate electromagnets, one each for vertical and horizontal deflection. These two electromagnet coils are constructed inside of a single assembly called a *deflection yoke*. The yoke can create the deflection because magnetic fields will bend the path of the electron beam emitted by the gun assembly.

The inductance of the yoke coils limits magnetic deflection to low-frequency applications. A TV receiver, for example, creates a lighted area on the screen called a *raster* by sweeping the vertical at approximately 60 Hz and the horizontal at 15,734 Hz. (*Note:* 15,750 Hz was used in the days before color TV, so some texts still use that figure—erroneously.) Medical oscilloscopes deal with very slow signals (up to either 100 Hz or 1000 Hz, depending upon the application), so often use magnetic deflection in order to take advantage of the shorter CRT length that is possible. But all service, engineering laboratory, and most scientific oscilloscopes use electrostatic deflection.

The CRT shown in Fig. 8-1(a) was an electrostatic type; note the vertical and horizontal deflection plates located between the second anode and the fluorescent screen. This type of CRT uses the electrostatic field between two plates to deflect the electron beam.

Figure 8-4 shows the position of the beam on the viewing screen as a function of the potentials applied to the deflection plates. When the potential on all four plates is the same, or zero as in Fig. 8-4(a), then the beam will be centered on the screen.

For the sake of this discussion let us assume the following notation for the four plates:

V_u is the upper deflection plate (vertical).
V_L is the lower deflection plate (vertical).
H_L is the left deflection plate (horizontal).
H_R is the right deflection plate (horizontal).

With these designations in mind, let us consider the following situations:

1. Figure 8-4(b). H_R is more positive than H_L; V_u and V_L are at zero volts. Beam moves to the right.

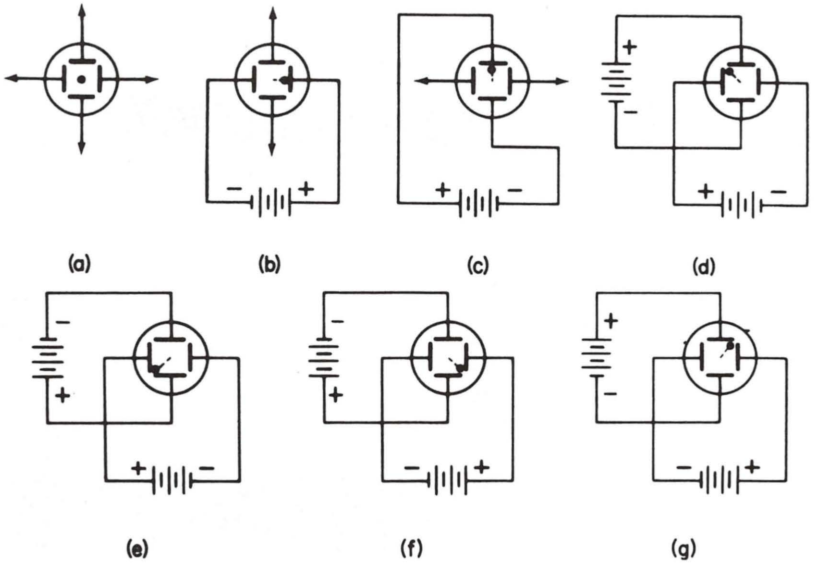


Figure 8-4. Position of the beam with different voltage situations at the deflection plates.

2. Figure 8-4(c). V_u is more positive than V_L; H_L and H_R are at zero volts. Beam moves upwards.
3. Figure 8-4(d). V_u is more positive than V_L; H_L is more positive than H_R. Beam moves diagonally to the upper left-hand corner of the screen.
4. Figure 8-4(e). V_L is more positive than V_u; H_L is more positive than H_R. Beam moves diagonally to the lower left-hand corner of the screen.
5. Figure 8-4(f). V_L is more positive than V_u; H_R is more positive than H_L. The beam moves diagonally to the lower right-hand corner of the screen.
6. Figure 8-4(g). V_u is more positive than V_L; H_R is more positive than H_L. The beam moves diagonally to the upper right-hand corner.

8-6 THE X-Y OSCILLOSCOPE

An X-Y oscilloscope (Fig. 8-5) allows the operator to position the beam anywhere on the CRT screen by controlling the potentials applied to the vertical and horizontal deflection plates. Although it is theoretically possible to use the deflection plates directly, in most commercial oscilloscopes vertical and horizontal amplifiers are used to drive the deflection plates. This arrangement allows us to display very low amplitude signals.

Note that most service and laboratory oscilloscopes sweep the horizontal plates with a time base signal, and so are "Y-time" (Y-T) designs. Many of

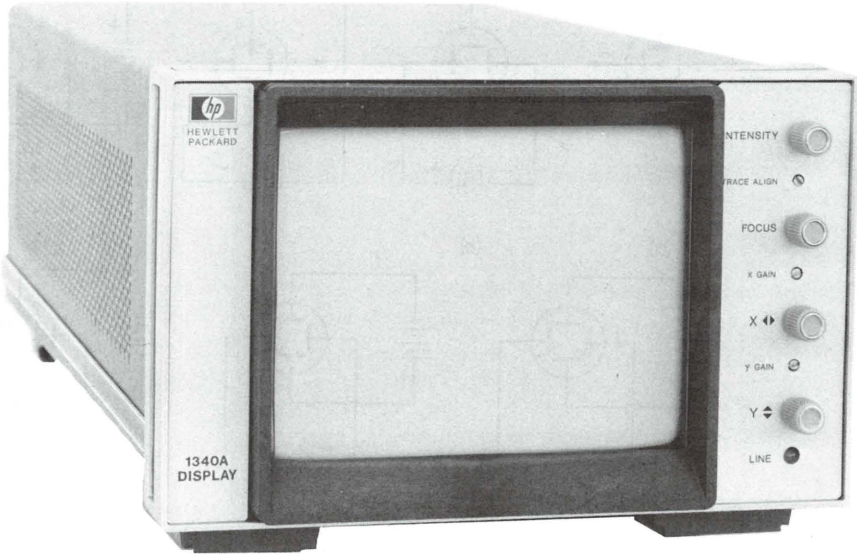


Figure 8-5. X-Y display oscilloscope. (Courtesy of Hewlett-Packard.)

those models (see Fig. 8-7) can be used in the X-Y mode because a selector switch allows the user to apply either the internal time base signal or an external signal to the horizontal deflection plates.

The X-Y oscilloscope is used in a wide variety of applications, especially in the fields of computer terminal display or scientific and medical instrumentation. It is very often the case that Tektronix or Hewlett-Packard X-Y oscilloscopes will be found as the readout display device in equipment produced by other manufacturers who find it less costly to buy an "O.E.M." (original equipment manufacturer) model than it is to design and build their own model.

Figures 8-6 and 8-7 show two applications for the X-Y oscilloscope. In Fig. 8-6 we see how to measure the relative *phase* between two signals of the same frequency, while in Fig. 8-7 we see the measurement of the frequency of two harmonically related signals. Both cases are examples of *Lissajous figures*.

Note in Fig. 8-6 that all phase differences other than 0, 90, 270, and 360 degrees produce elliptical patterns. The tilt direction of the ellipse limits the possible phase difference between V_x and V_y to two quadrants, while the "fatness" of the ellipse gives us the actual phase. Using the alphabetical notation of Fig. 8-6, we can calculate the phase angle from the expression

$$\Theta = \arcsin (a/b) \tag{8-1}$$

There are two special cases in Fig. 8-6: 0 deg and 90 deg. When both vertical and horizontal plates are connected to the same signal, or to in-phase signals of the same frequency, then the phase difference is 0 deg or 360 deg (same actual point). This phase difference produces a straight line on the CRT screen. If both signals are of the same amplitude, then the line will be at a 45-deg angle from vertical or horizontal lines. But if the signal amplitudes are different, then the tilt angle will be

$$\Theta_t = \arctan \left[\frac{E_v}{E_h} \right] \tag{8-2}$$

where Θ_t = the tilt angle in degrees
E_v = the vertical potential
E_h = the horizontal potential

Example 8-1

Calculate the tilt angle if in-phase 60-Hz ac signals are applied to the deflection plates so that E_v is 38 volts and E_h is 14 volts peak.

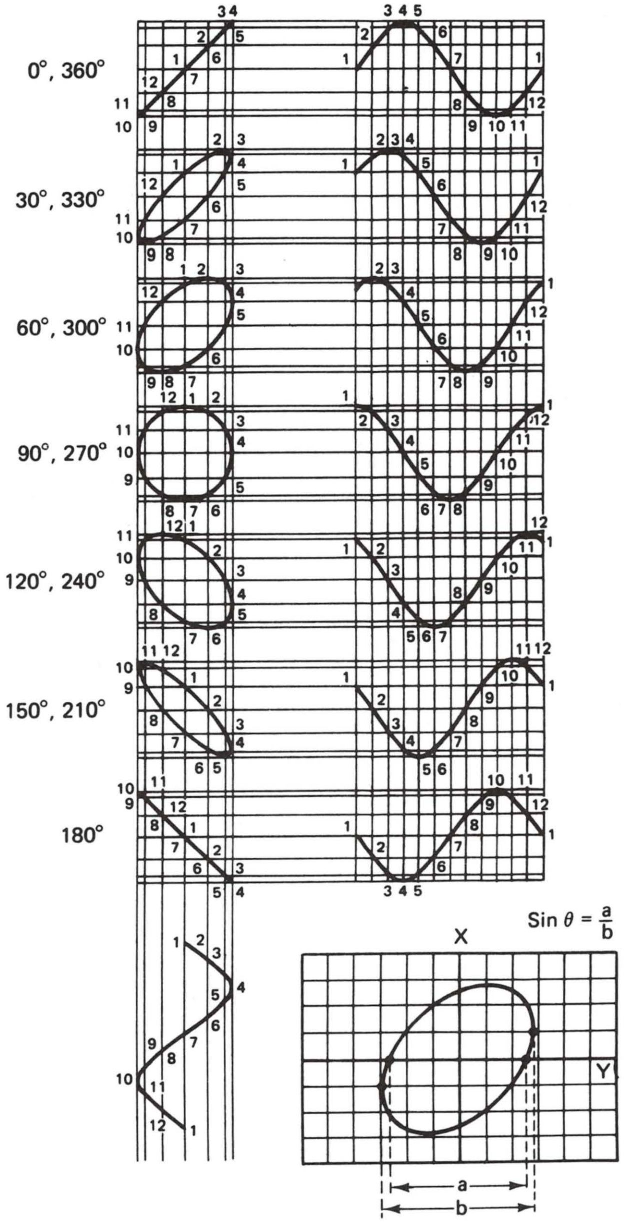


Figure 8-6. Lissajous patterns on X-Y oscilloscope with signals of same frequency but differing phases.

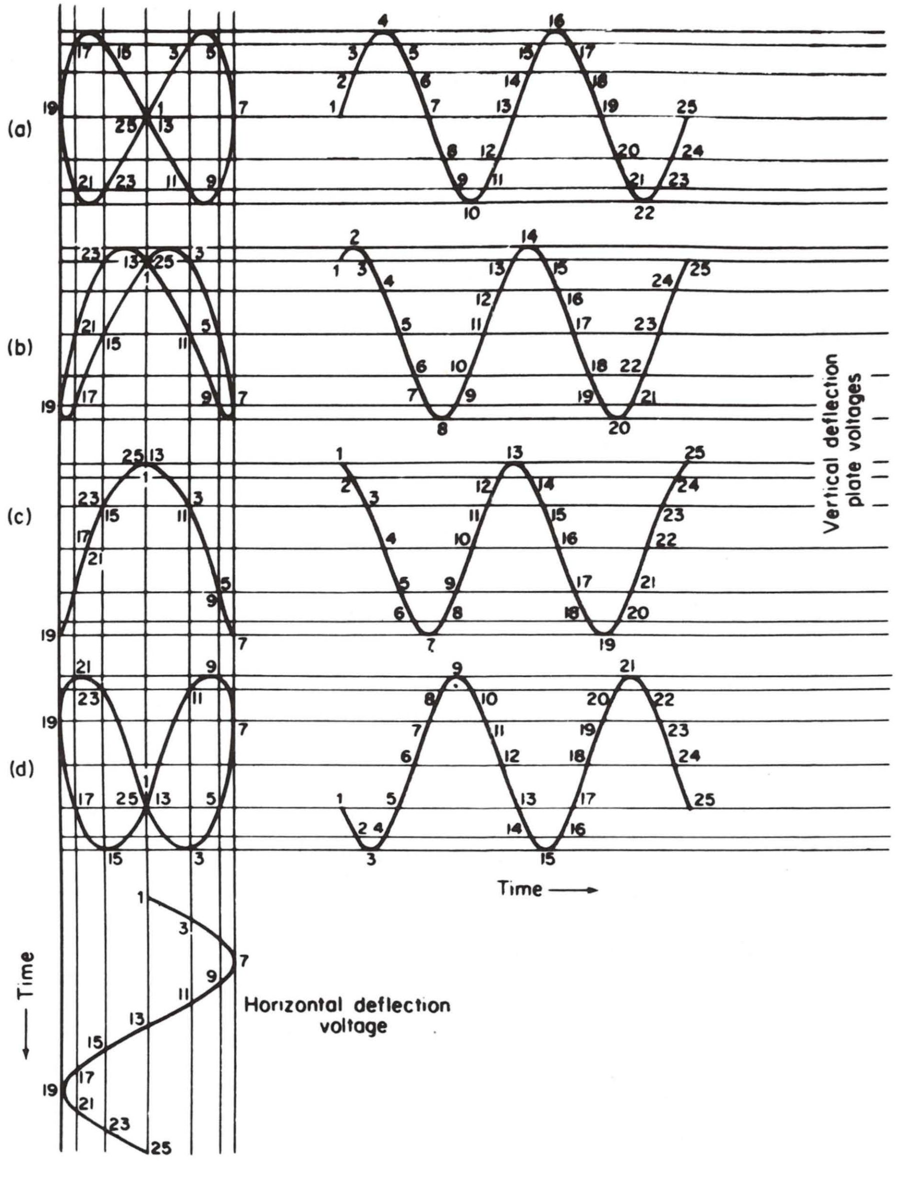

(a)
(b)
(c)
(d)
Time
Vertical deflection plate voltages
Time
Horizontal deflection voltage

Figure 8-7. Lissajous figures for 2:1 frequency ratio.

SOLUTION

$$\Theta_t = \arctan (E_v/E_h) \qquad (8\text{-}2)$$
$$\Theta_t = \arctan (38/14)$$
$$\Theta_t = \arctan (2.71) = \mathbf{69.8\ degrees}$$

The Lissajous patterns of Fig. 8-7 result from applying harmonically related signals to the vertical and horizontal deflection plates. In the example shown, the vertical frequency is twice the horizontal frequency. Note in Fig. 8-7 that the *phase difference* determines the actual *shape* of the Lissajous pattern, while the *number of loops* is determined by the *frequency ratio*. The two frequencies are related by

$$\frac{F_v}{F_h} = \frac{N_h}{N_v} \qquad (8\text{-}3)$$

where F_v = the frequency applied to the vertical plates
F_h = the frequency applied to the horizontal plates
N_v = the number of loops observed along the vertical edge
N_h = the number of loops observed along the horizontal edge

Example 8-2

In Fig. 8-7(a) we know that the vertical signal is precisely 1 kHz. Find the frequency of the horizontal signal.

SOLUTION

$$F_v/F_h = N_h/N_v \qquad (8\text{-}3)$$

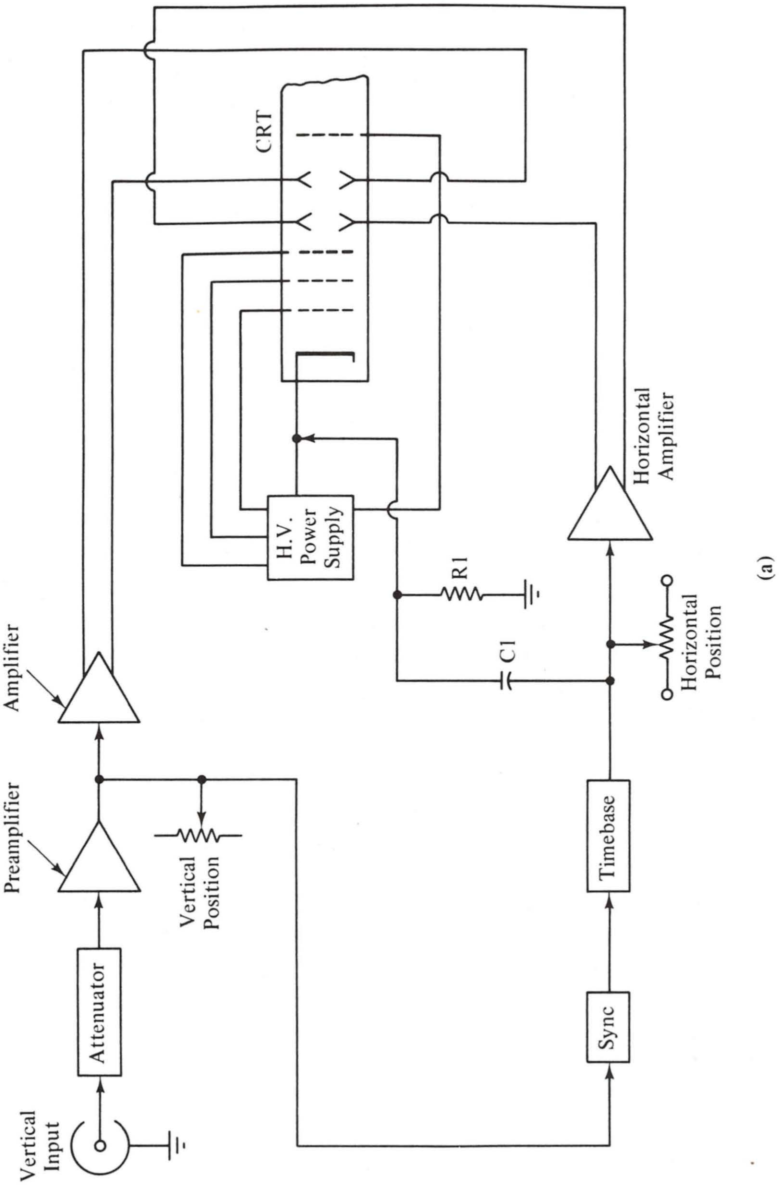
so

$$F_h = F_v N_v/N_h$$
$$F_h = (1000\ \text{Hz})(1\ \text{loop})/(2\ \text{loops})$$
$$F_h = 1000\ \text{Hz}/2 = \mathbf{500\ Hz}$$

This technique is limited by the necessity of accurately knowing one of the frequencies, and it must be an integer multiple or submultiple of the unknown. In practical applications a variable, calibrated, signal source is adjusted until the pattern "locks in," and then the frequency is read from the generator dial.

8-7 THE Y-T OSCILLOSCOPE

The instrument that many people identify as the *basic* oscilloscope is the Y-time, or Y-T, oscilloscope shown in Fig. 8-8(a). The vertical channel is the same as in



(a)

Figure 8-8. (a) Block diagram for a Y-time oscilloscope.

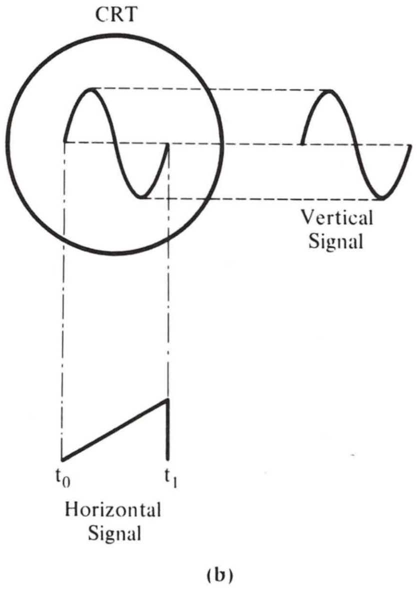


Figure 8-8. *(cont.)* (b) Relationship of vertical to horizontal signals produces the pattern on the screen.

the X-Y system; push-pull output vertical signal amplifiers driving the deflection plates. The horizontal amplifier, however, does not connect to an external signal source, but to an internal time base signal.

The time base generator creates a sawtooth waveform that deflects the beam horizontally. The CRT deflection plate potentials are arranged so that the beam is positioned on the left side of the screen when the sawtooth ramp voltage is zero. The beam is pulled to the right as the ramp voltage rises. If everything is adjusted properly, then the ramp will reach maximum just as the beam disappears off the right edge of the screen.

The sawtooth waveform drops rapidly back to zero once the ramp portion is completed, snapping the beam back to the left edge of the CRT screen. This action would result in a *retrace line*'s being printed on the CRT screen. To overcome this problem, a *retrace blanking pulse* is generated that extinguishes the electron beam. This turns off the beam, eliminating the retrace line on the screen.

In low-cost oscilloscopes the time base is said to be *free-running,* although the time base oscillator may, in fact, be synchronized to the vertical amplifier signal. Unless the time base is so synchronized, the waveform marches across the screen and remains unstable. Synchronization means that the time base signal sweeps across the screen in a time that is equal to an integer number of vertical waveform periods. The vertical waveform will then appear locked on the CRT screen.

The vertical section consists of a wideband preamplifier and power amplifier combination that drives the CRT vertical deflection plates. The vertical amplifier has a high gain, so large signals must be passed through an *attenuator,* or, in low-cost instruments, a *vertical gain* control.

The free-running oscilloscope is a low-cost instrument, but it is relatively simple to operate. An example is shown in Fig. 8-9. The four controls surrounding the CRT screen are part of the high voltage power supply.

The vertical amplifier is controlled by a two-part attenuator, labeled here as *vertical gain* and *vertical attenuator,* although these are actually only fine and coarse attenuators, respectively.

The horizontal section is controlled by a *horizontal gain* control and two sweep frequency controls: *sweep selector* and *sweep vernier*. The horizontal gain control determines the width of the sweep when the oscilloscope's internal time base is used, and the size of the horizontal signal when the horizontal amplifier is connected to the external *horizontal input*.

The horizontal sweep frequency is determined by two controls. The *sweep selector* is a coarse frequency (or range) switch. It will, for example, select ranges of 20 to 120 Hz, 120 to 1400 Hz, etc. The sweep vernier, on the other hand, "fine tunes" the horizontal frequency to the specific point desired within the range. The sweep selector also selects *external sweep* (connects the horizontal sweep amplifiers to the horizontal input connector) and *line frequency* (connects the horizontal amplifier to a sample of the 60-Hz ac power mains signal).

The synchronization function is controlled by a *sync lock* control and a *sync selector* switch. The selector switch determines the source of the sync signal: external, ac line frequency, or internal. In the latter case, the sync is taken from the vertical amplifiers.

Two further controls that are of interest are the *line phasing* and *1 V P-P screen cal*. The line phasing control varies the phase of the 60-Hz sample (usually taken from a filament transformer) used in the line frequency positions of the sweep selector and sync switches.

8-8 TRIGGERED SWEEP

The low-cost free-running oscilloscope provides only limited usefulness. The triggered sweep oscilloscope, on the other hand, is considered far more versatile, and is the standard of the industry. Tektronix, Inc. and others now offer low- to medium-priced triggered sweep models that have supplanted the free-running models in all but the lowest cost applications.

Figure 8-10 shows the block diagram of a triggered sweep model, at least of those stages that differ from free-running models. In a triggered sweep model, the horizontal sweep is turned off, with the CRT beam remaining at the left-hand side of the CRT screen, until a vertical signal is encountered.

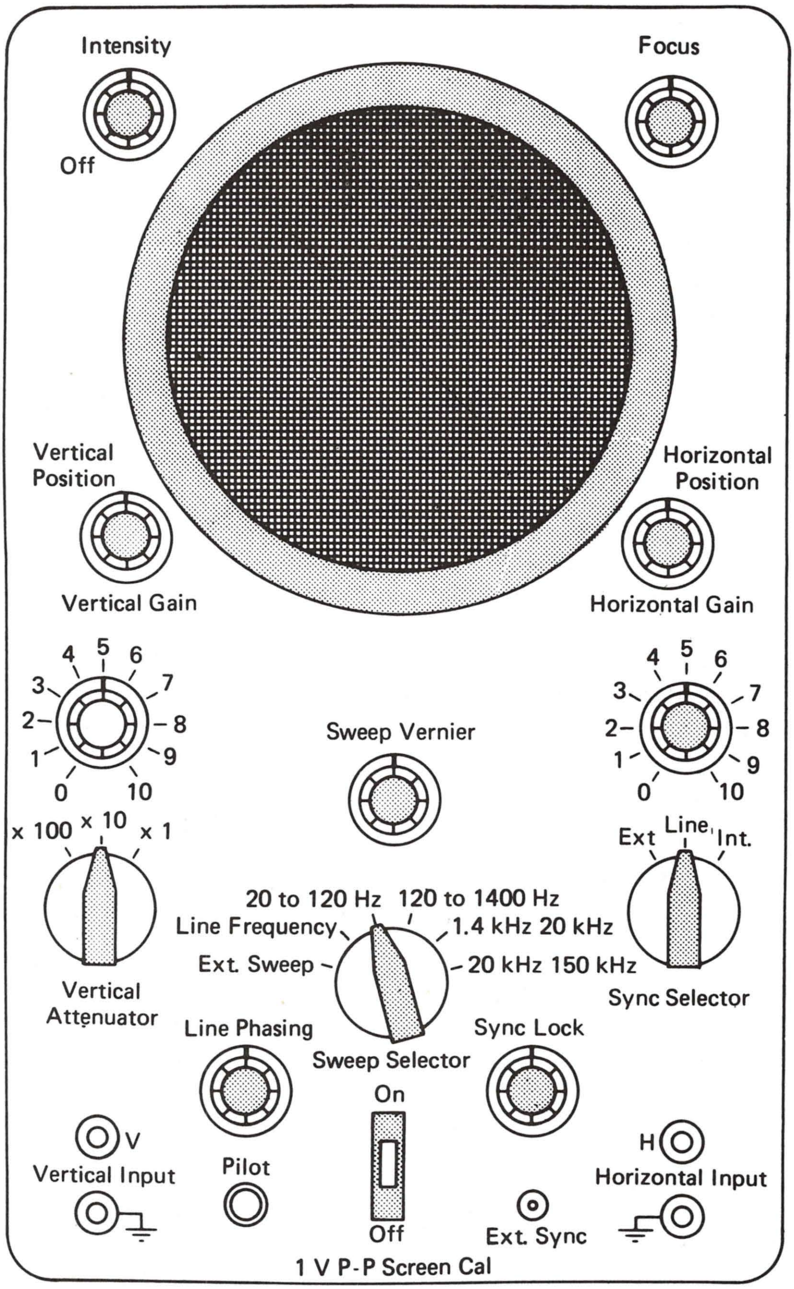


Figure 8-9. Front panel of a low-cost Y-time oscilloscope.

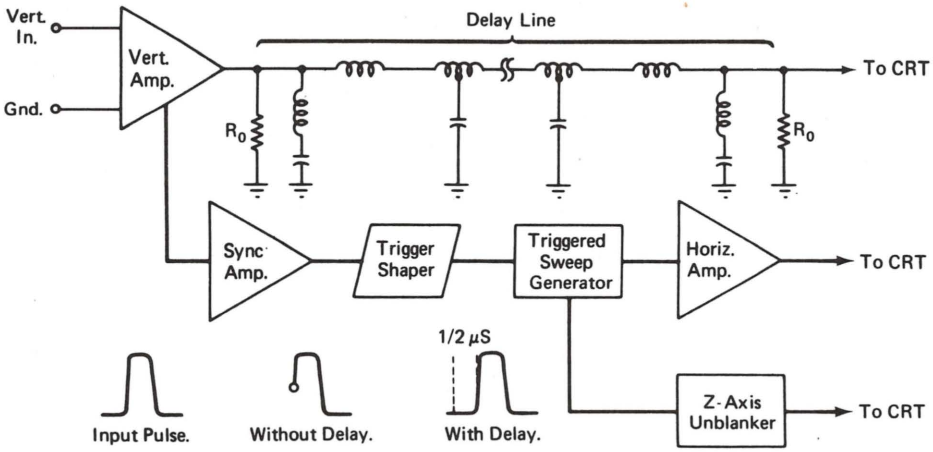


Figure 8-10. Block diagram of a triggered sweep oscilloscope.

The trigger circuit usually consists of a Schmitt trigger and a monostable (i.e., one-shot) multivibrator. The Schmitt trigger is a circuit that will produce either high or low output levels, depending upon whether or not the input signal is above or below a present threshold level. When the amplitude of the vertical amplifier signal exceeds this threshold, then the trigger output changes level, and this action causes the one-shot to produce a pulse that unlocks the sweep for one excursion across the CRT screen. As long as signals exist in the vertical amplifier and these have a level that is capable of retriggering the sweep, then the sweep will have the appearance of being free-running. The difference, however, is that the sweep begins at the same point on the vertical waveform each time.

All electronic circuits have a certain propagation time, that is to say the length of time delay, or phase shift, between the input and the output. The trigger circuit is no exception; it takes time for these circuits to operate. If the vertical signal were applied directly to the CRT deflection plates, then part of the vertical waveform leading edge would be lost; the sweep would not begin until several nanoseconds into the waveform. To overcome this problem, manufacturers place a *delay line* between the output of the vertical amplifier and the CRT. The delay lines in most oscilloscopes have a fixed delay time, but in a few expensive models, the delay line is made variable to accommodate signals of various rise times.

This delay line should not be confused with *delayed sweep,* or *delayed triggering,* found in some models. The delayed trigger feature creates a small time delay between the beginning of a waveform and the initiation of a sweep cycle. The delayed sweep feature allows viewing of a small feature, or section, of a larger waveform—for example, a small oscillation, or some ringing, that is riding on a larger waveform. We may, for example, see several cycles of a

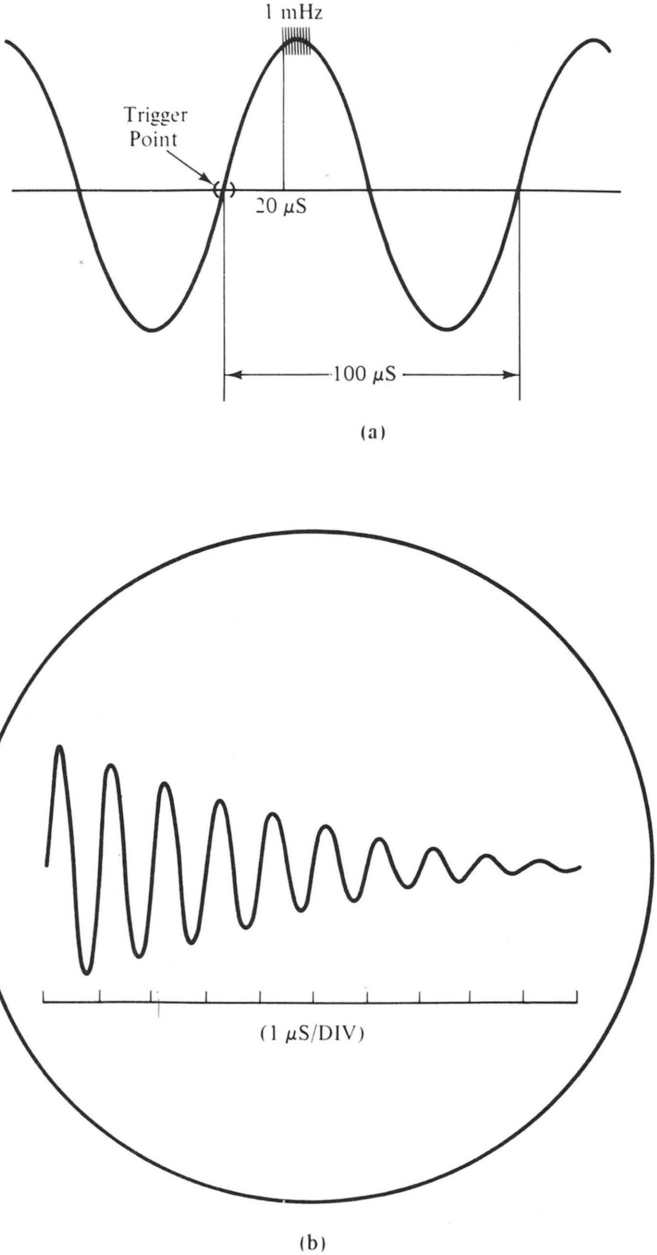


Figure 8-11. (a) Input signal with a higher frequency oscillation on its peak. (b) Using delayed sweep allows us to see the high frequency signal.

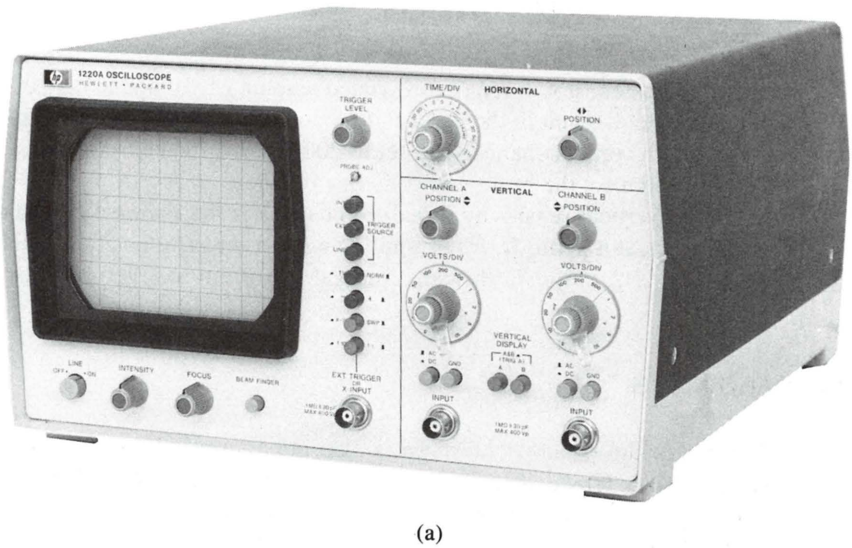
1-mHz oscillation riding on the peak of a 10-kHz sine wave [see Fig. 8-11(a)] The period of a 1-mHz signal is 1 μs, while that of a 10-kHz signal is 100 μs. In order to view the 1-mHz signal, set the horizontal sweep to show several cycles of the 1-mHz waveform, and then trigger the sweep on the 10-kHz signal. It can be observed in Fig. 8-11(a) that the 1-mHz oscillation begins at approximately 20 μs past the trigger point. If the *trigger delay* control is adjusted for 20 μs, and the sweep set for 1 μs/div, we would see only the 1-mHz signal on the CRT screen. The 10-kHz signal would be suppressed.

8-9 OSCILLOSCOPE SPECIFICATIONS

There are several parameters that are important to the user of oscilloscopes. How well these are understood may determine how useful the instrument is in any given situation.

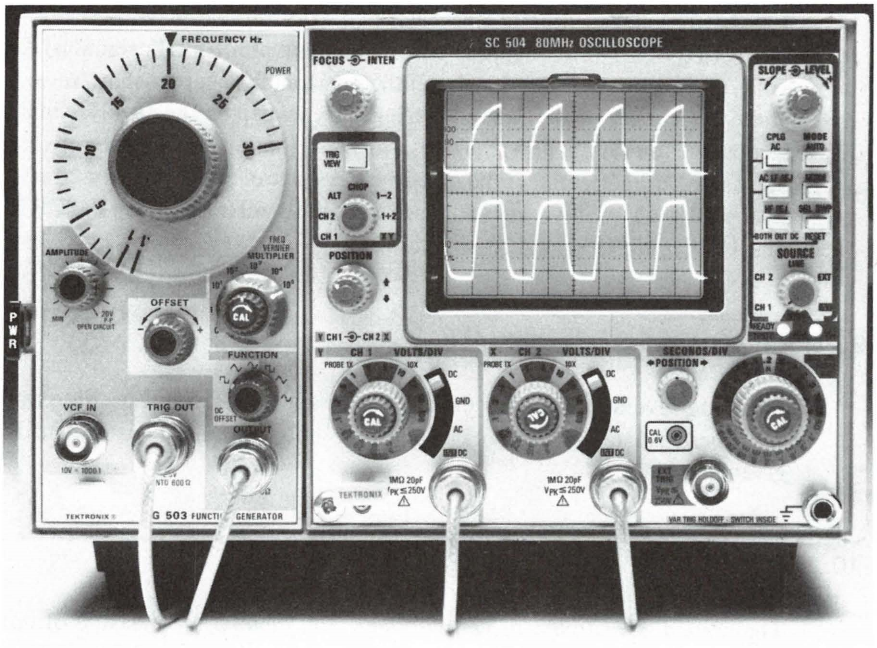
8-10 SENSITIVITY

The *vertical sensitivity,* or *vertical deflection factor,* is a measure of how much deflection will be given for a specified input signal. The CRT screen has a grid pattern called a *graticule* (see Fig. 8-12). The divisions of the graticule are between 0.75 and 1.3 cm apart, with most being exactly 1 cm apart. The vertical



(a)

Figure 8-12. (a) Portable triggered sweep oscilloscope. (Courtesy of Hewlett-Packard.)



(b)

Figure 8-12. *(cont.)* (b) Small dual-trace oscilloscope part of a Tektronix TM-500 mainframe. (Courtesy of Tektronix, Inc.)

attenuator(s) are calibrated in terms of *volts per division,* meaning that, at the setting indicated, it will require the stated amount of voltage to deflect the beam one graticule division in the vertical direction. In the oscilloscope shown in Fig. 8-12(a) both vertical channels are set to 200 mV/div. This means that a 200-mV signal will deflect the beam vertically one division.

The vertical sensitivity of an oscilloscope is the *smallest* deflection factor on the attenuator knob. If, for example, the most sensitive position of the vertical attenuator is 2 mV/div, then the vertical sensitivity of that oscilloscope is 2 mV/div.

8-11 BANDWIDTH

The bandwidth of an oscilloscope is also important, but comes in for a lot of, shall we say, "creative spec writing," in advertisements. A typical oscilloscope may have a frequency response from dc to some specified upper frequency limit (e.g., 500 kHz, 15 mHz, 35 mHz, etc.). Most commonly, reputable oscilloscope manufacturers specify the range of frequencies over which the vertical amplifier gain (hence attenuator calibration) is within ±3 dB of the proper gain. Beware of oscilloscopes that have a frequency response specified to some other point, i.e., ±3.5 dB. This is not the same, and comparisons are difficult to make.

Also beware of those models that offer a response specification of "3 dB down at 10 mHz," or some other frequency. This would lead you to believe that the bandwidth curve is flat from dc to some high frequency, and that it falls off to its rated frequency only 3 dB. You may easily believe that this is a model with a 10-mHz bandwidth. But it could also be true that the gain rolls off much faster than implied by the specification, and that the manufacturer "tweaked" the response to improve the gain at 10 mHz, in order to validate the specification. A reputable manufacturer guarantees the specification across the entire band.

Often more important than basic frequency response is the *rise time* of the vertical amplifiers. This specification gives us an indication of how well the oscilloscope will reproduce pulse waveforms.

8-12 RISE TIME

The rise time of a pulse is usually defined as the time required for its leading edge to rise from 10 percent of its final amplitude to 90 percent.

An oscilloscope used to view a pulse needs a vertical rise time that is equal to or faster than the pulse rise time. If the oscilloscope manufacturer does not publish the rise time specification, then it may be calculated from the frequency response by using Eq. (8-4). Alternatively, the same equation may also be used to calculate the required bandwidth if the pulse rise time is known.

$$F = \frac{0.35}{t_r} \tag{8-4}$$

where F = the ±3-dB frequency response in hertz (Hz)
t_r = the pulse rise time in seconds (s)

Example 8-3

What is the minimum frequency response required of an oscilloscope that is to reproduce without distortion a pulse that has a 15-nanosecond (1.5×10^{-8} s) rise time?

SOLUTION

$$F = 0.35/t_r \tag{8-4}$$
$$F = 0.35/1.5 \times 10^{-8}\ \text{s}$$
$$F = 2.3 \times 10^{7}\ \text{Hz} = \mathbf{23\ mHz}$$

8-13 HORIZONTAL SWEEP TIME

The low-priced free-running oscilloscope uses a horizontal section that is calibrated in sweep frequency. Better models, including all of those used in engineering, calibrate the horizontal section in units of *time per division* [see Figs. 8-12(a) and 8-12(b)]. The Tektronix model shown in Fig. 8-12(b) is set to sweep at 0.2 μs/div. Since these are models that have ten horizontal graticule divisions, the sweep time across the CRT screen is

$$\frac{0.2\ \mu\text{s}}{\text{div}} \times 10\ \text{div} = 2\ \mu\text{s}$$

Using this information, we may determine the approximate frequency or period of displayed waveforms. In Fig. 8-12(b), for example, the bottom trace shows that each complete pulse has a *period* of 2.3 divisions, which in units of time converts to

$$2.3\ \text{div} \times \frac{0.2\ \mu\text{s}}{\text{div}} = 0.46\ \mu\text{s}$$

The frequency in hertz is the reciprocal of the period in seconds, so we determine that the frequency of the displayed waveform is

$$F_{\text{Hz}} = \frac{1}{\left[0.46\ \mu\text{s} \times \dfrac{1}{10^6\ \mu\text{s}}\right]}$$

$$F_{\text{Hz}} = \frac{10^6}{0.46\ \text{s}}$$

$$F_{\text{Hz}} = 2.2 \times 10^6\ \text{Hz} = 2.2\ \text{mHz}$$

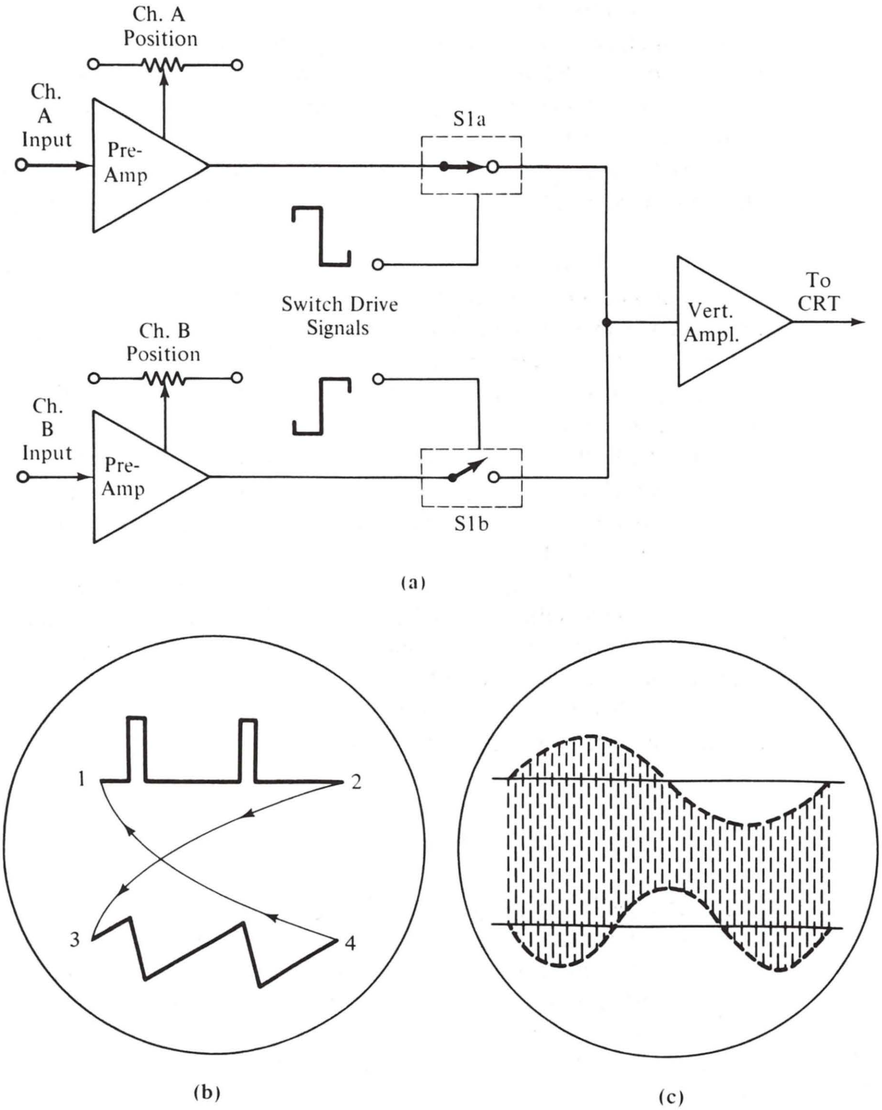
The oscilloscope is uniquely suited to waveform analysis, not merely giving shape information, but also frequency, period, amplitude, and rise time.

8-14 DUAL-BEAM MODELS

The popularity of the dual-beam oscilloscope has grown immensely in the past decade or so. The name "dual beam," however, does not refer to two electron beams, but to two *traces* on the CRT screen. The creation of the two traces from a single electron beam is done by switching.

The oscilloscopes in Fig. 8-12 are both dual-beam models, so they allow the user to view and compare two harmonically related waveforms. Note that these are not like older two-channel models that allowed you to view *either* channel A *or* channel B, but not *both* at the same time.

Figure 8-13 shows how a dual-beam oscilloscope can be made. There is a



(b)

(c)

Figure 8-13. (a) Electronic switching allows two traces on the same CRT. (b) ALTERNATE mode. (c) CHOP mode.

single vertical output amplifier driving the vertical deflection plates of the CRT. But there are two preamplifiers, one each for channels A and B.

A dual CMOS electronic switch is used to switch the input of the vertical amplifier between the outputs of the two vertical preamplifiers. These switches are driven *out of phase* with each other, so that only one is connected to the vertical amplifier at any given time.

The oscilloscope in Fig. 8-12(b) calls the two channels "1 and 2," and there are six modes in which they can be operated (see the switch inside the block to the left of the CRT screen): channel 1, channel 2, ALT, CHOP, 1 − 2, and 1 + 2. The "X-Y" refers to using an external horizontal signal instead of the internal time base signal.

The "Ch. 1" position means that only channel one will be displayed on the CRT screen. Similarly, "Ch. 2" means that only the channel two signal will be displayed.

The ALT and CHOP positions refer to *alternate* and *chopped* two-trace displays in which both channels will be displayed. There is, however, a critical difference between these modes. In the ALT mode, the vertical amplifier will be connected first to channel 1 for one entire sweep, and then to channel 2 during the next sweep. That is to say that channels one and two are displayed one at a time on alternate sweeps.

In the CHOP mode (that is, the type of circuit shown in Fig. 8-13 is operating) both waveforms are displayed during each sweep. The frequency of the signal driving the beam switching circuit must be much higher than the frequency of the signal being displayed.

The "1 − 2" and "1 + 2" positions are single-beam positions, but use both vertical preamplifier channels. The "1 − 2" position displays the *difference* between two input signals, whereas the "1 + 2" position displays their *sum*. Note that most oscilloscopes offer only the difference function, i.e., "1 − 2" or "A − B." Also, the vertical deflection factor is calibrated *only* if the input attenuators for both channels are set to the *same* volts/div factor.

8-15 STORAGE OSCILLOSCOPES

The input signal must be periodic, that is to say, it must repeat itself at fixed intervals, before the pattern will remain stable long enough to be analyzed. But single events often pass too rapidly to be viewed on an oscilloscope unless it is equipped with a *storage* feature. There are now two basic types of storage oscilloscope, one type using a special CRT and the other using digital memory techniques. Figure 8-14 shows the block diagram of the type that requires a special storage CRT.

A storage oscilloscope retains the image of the waveform on the screen for a period of time before it fades out. In the class of instruments using special CRT

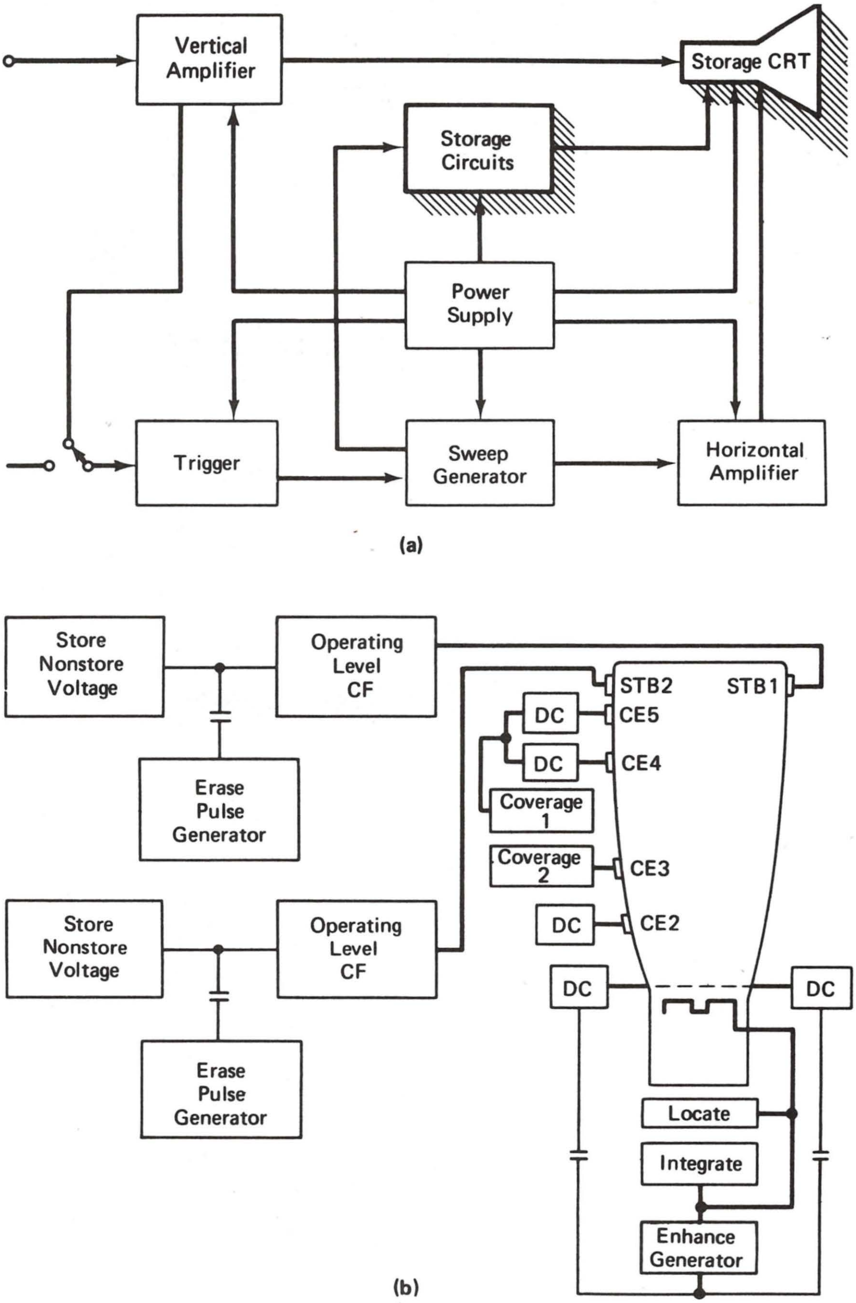


Figure 8-14. Functions of typical storage-type oscilloscopes. (a) Block diagram of typical storage scope. (b) Subdivision of the storage circuits.

designs, there are several subclasses that operate on slightly different principles. Figure 8-15 shows three types of storage CRT.

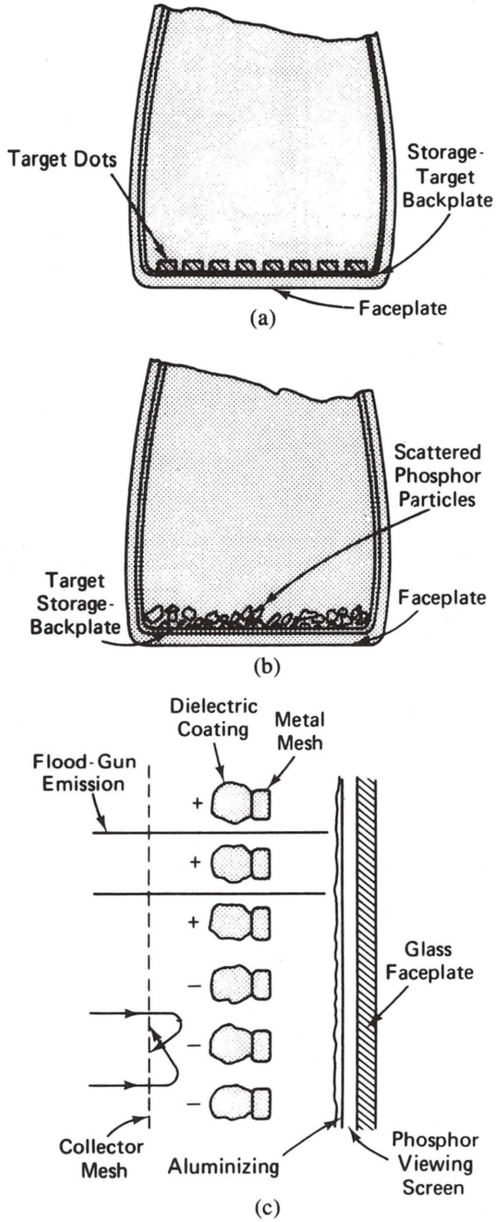


Figure 8-15. Three types of storage CRT.

The two types of CRT shown in Figs. 8-15(a) and 8-15(b) depend for operation on a phosphor screen in which individual particles are insulated from each other. In the case of Fig. 8-15(a) the phosphors lie in the same plane to form target dots, while the other type [Fig. 8-15(b)] uses layers of scattered particles of increasing weights. A special *flood gun* in the CRT emits high-energy electrons that excite the phosphors. The phosphor particles struck by these electrons take on a charge of 150 to 200 volts, but unenergized particles remain at zero volts. When electrons from the main electron gun have preenergized certain phosphors, forming an image on the CRT screen, these phosphors attract more flood gun electrons. The image is thereby retained by the continuous bombardment of flood gun electrons. Erasure of the screen is accomplished by grounding the phosphor screen, thus returning all phosphor particles to a zero volts potential.

The other type of special CRT is the wire-mesh variety shown in Fig. 8-15(c). The flood gun charges one mesh so that no further electrons can pass through, although the *write* or *main gun* electrons will pass through if sufficiently energetic.

A *split-screen* storage oscilloscope allows the operator to store waveforms in either the top or bottom half of the screen. If both halves are turned on, then the device will store traces appearing anywhere on the screen.

Some storage oscilloscopes have a feature called *variable persistence,* which allows the operator to vary the length of time that the image will remain on the screen.

8-16 DIGITAL STORAGE

A digital storage oscilloscope (see Fig. 8-16) uses an analog-to-digital converter to *digitize* the input waveform. That means it will sample the waveform at many points, and then convert the instantaneous amplitude at each point to a binary number value proportional to the amplitude. These binary numbers are then

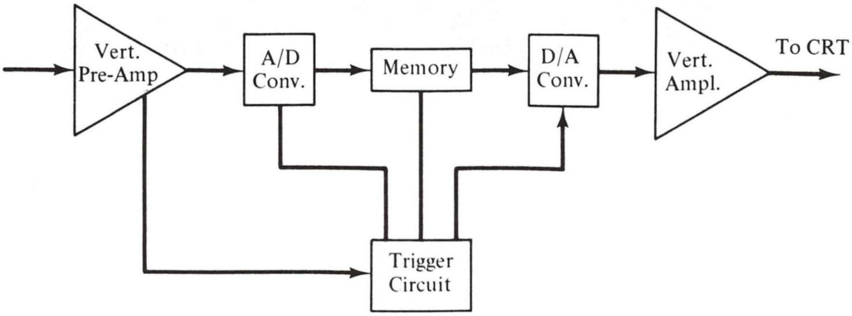


Figure 8-16. Digital storage oscilloscope.

stored in memory. A digital-to-analog converter at the output of the memory circuit reconverts the binary words to analog voltages capable of driving the CRT vertical deflection system. The memory is scanned many times per second, so the CRT screen is constantly being "refreshed" by the data stored in memory before it can fade out.

The digital storage technique is used in some high-frequency oscilloscopes designed as "transient catchers," but the technique is very expensive for this frequency range. There is a sampling rule called Nyquist's theorem that requires a sampling rate for any given waveform of *twice* the highest frequency Fourier series component making up the waveform. Consider a short, irregularly shaped pulse. It may contain Fourier series components up to 10 mHz or so. To accurately sample that pulse would require 2×10^7 samples per second. Even if the pulse is very short, a lot of memory is used, and memory is expensive—as are ADC and DAC circuits that will track that fast.

The digital storage technique does, however, find extensive use in medical and physiological oscilloscopes, such as that shown in Fig. 8-17(a). The sweep time is typically 25 mm/s, so it takes approximately 4 seconds to sweep the CRT,

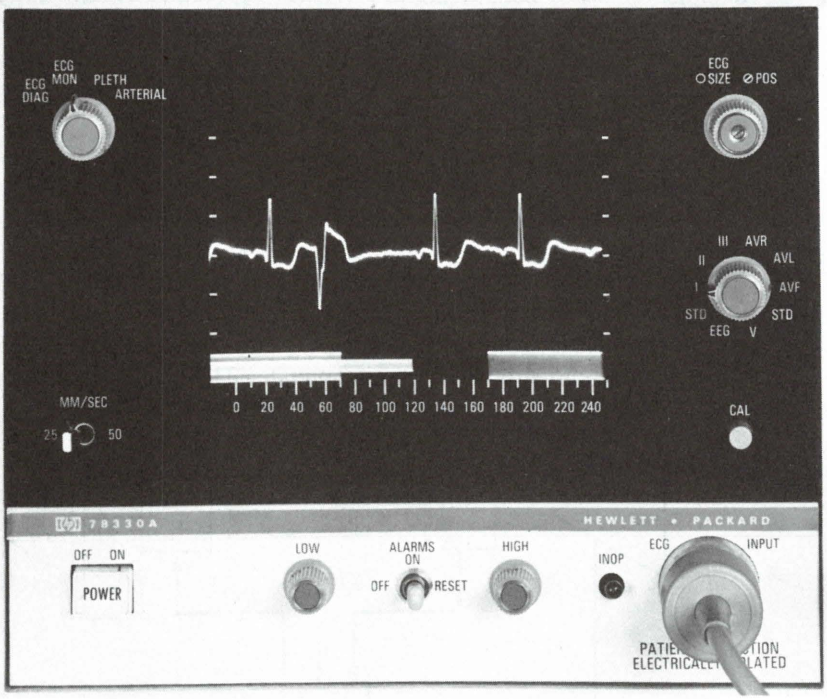


Figure 8-17(a). Medical "memory 'scope" uses digital technique. (Courtesy of Hewlett-Packard.)

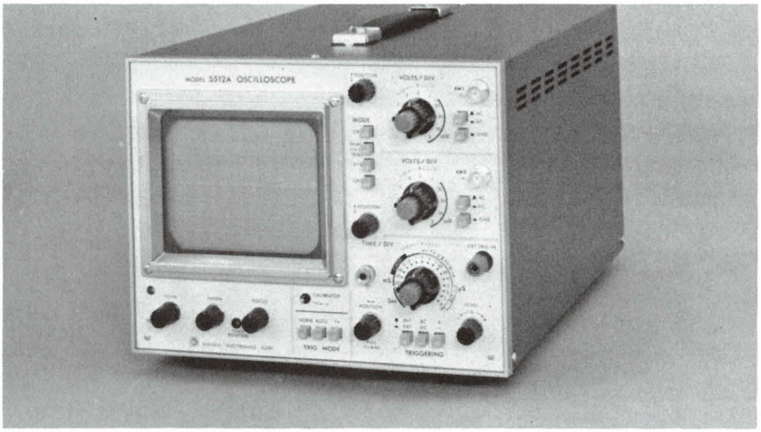


Figure 8-17(b). Storage oscilloscope. (Courtesy of Kikasui)

so one would normally expect to find the beam fading on the left side of the screen before the trace reaches the right side. Because of the digital storage capability, however, this 'scope is a *nonfade* model. The waveform on the screen is refreshed continuously, until a new waveform overwrites the old data. Many of these models have a *freeze* control that turns off the ADC so that no new data are written into memory. This action leaves the old waveform on the screen to be refreshed indefinitely. A standard CRT storage oscilloscope made by Kikasui is shown in Fig. 8-17(b).

8-17 OPERATING THE TRIGGERED-SWEEP OSCILLOSCOPE

Learning to cope with the dual-beam, high-frequency, triggered-sweep oscilloscope is *fundamental* to an engineering or technology education. It is unfortunate that some schools do not emphasize this skill sufficiently, because you will not be able to function properly in industry or the sciences until it is mastered. Here we can give you some fundamental guidelines, but nothing, repeat nothing, written in book form will substitute for going to the laboratory and practicing with various types of input signal.

8-18 OSCILLOSCOPE CONTROLS

In this discussion we will cover most controls commonly found on modern oscilloscopes, but specific reference is made to the H-P model 182C shown in Fig. 8-18. This oscilloscope is an example of a main-frame model in which various

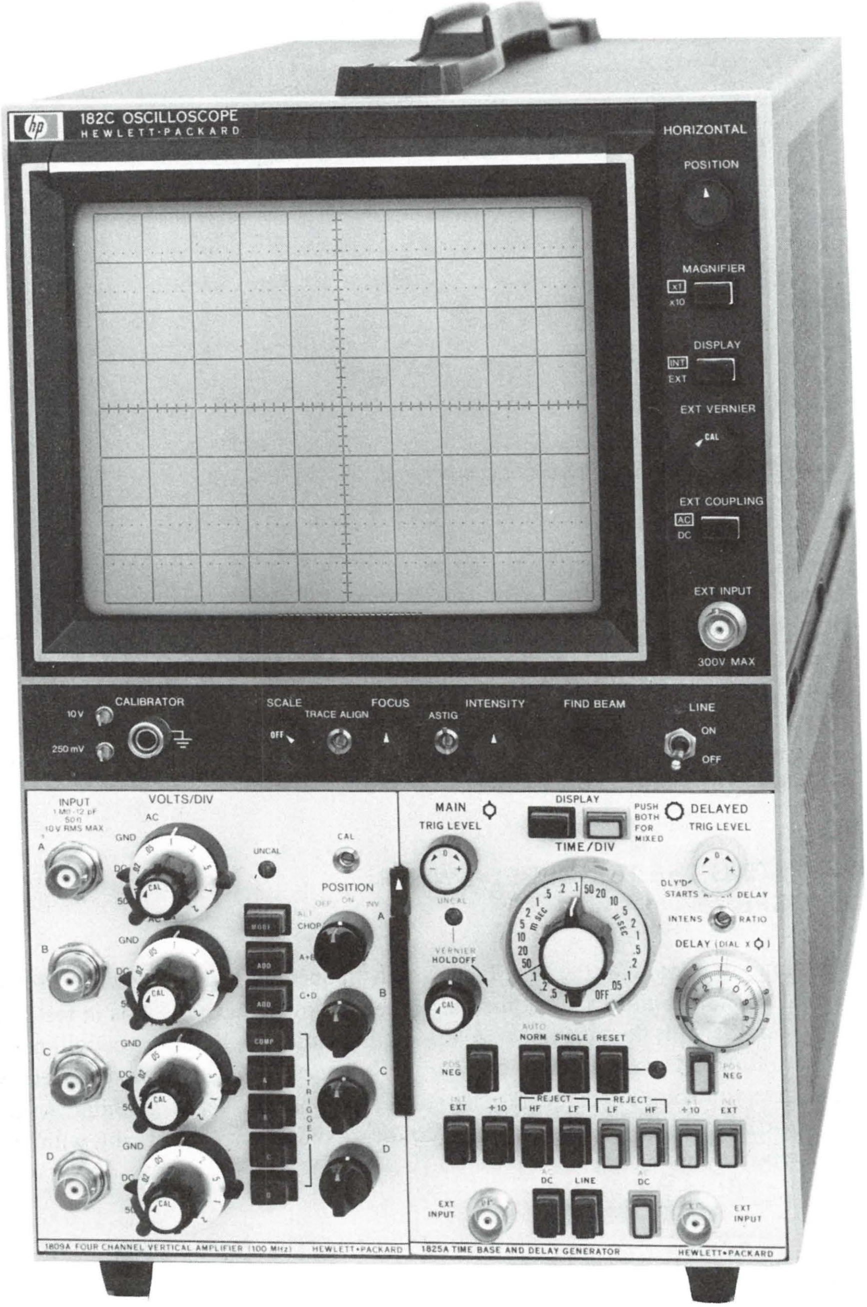


Figure 8-18. Main-frame oscilloscope. (Courtesy of Hewlett-Packard.)

vertical and horizontal plug-ins fit into the main frame containing the CRT, power supplies, and CRT control circuits. This type of oscilloscope can be custom configured to each user's requirements.

Horizontal controls in Fig. 8-18 are located on a horizontal time base plug-in (of which, more in a moment) and along the edge of the upper unit to the right of the CRT screen. The controls located in the latter place include *position, X1-X10 magnifier, display, EXT vernier, EXT coupling,* and *EXT input.* These controls function as follows:

> *Position.* Shifts the beam horizontally left and right across the CRT face. The most obvious effect of this control is that it shifts the starting point of each sweep, although in reality the whole trace moves. It can be used to move key features of a waveform directly over a graticule line for easier measurement of period.
>
> *X1-X10 magnifier.* This control amplifies the sweep speed. In the X1 position, the time base indications on the time/div switch (located on the time base plug-in below) are accurate. But in the X10 position, each setting of the time/div knob results in a sweep speed that is ten times faster than that indicated. Example, setting the time/div knob to 0.1 ms/div will yield the 0.1 ms/div speed if the magnifier is in the X1 position, and 0.01 ms/div if the magnifier is in the X10 position.
>
> *Display (INT/EXT).* Determines whether the horizontal amplifier is connected to the time base (INT position) or to the external input (EXT position).
>
> *EXT vernier.* Adjusts the width of the horizontal display when the *display* switch is in the EXT position.
>
> *EXT coupling (ac/dc).* Determines whether external input is ac coupled or dc coupled to the horizontal amplifier.
>
> *EXT input.* External input to the horizontal amplifier.

The remainder of the main-frame controls are located immediately below the CRT screen, and include: *calibrator, scale, trace align, focus, astig, intensity, find beam,* and *line on/off.*

> *Calibrator.* Produces a square wave (usually 1 kHz on most models) of either 250-m V or 10-V amplitude (other models use 1 V). Touching the vertical amplifier input probe to these points displays the calibration signal on the CRT screen.
>
> *Scale (on/off).* Illuminates, usually by side lighting, the graticule on the CRT screen to allow it to show up when the trace is photographed.
>
> *Trace align.* Screwdriver adjustment that compensates for the effects of local magnetic fields on the CRT trace. This control is adjusted so that the CRT beam is horizontal with respect to the graticule. It should be adjusted only when needed, and this usually occurs when the oscilloscope is moved to a different location.

Focus. Adjusts size and "roundness" of the CRT beam.

Astig (astigmatism). This control is used to insure the "roundness" of the beam across the entire width of the CRT. It is adjusted so that the horizontal sweep line is of uniform thickness across the entire width of the screen.

Intensity. Controls the brightness of the CRT beam.

Find beam. Disconnects vertical and horizontal inputs and displays a dot at the point where the position controls place the beam, regardless of front panel controls.

Line (on/off). AC mains power switch. Turns the instrument on and off.

The vertical plug-in is the unit located in the lower left corner (Fig. 8-18). This particular model has four vertical channels, labeled A through D. The input connectors are type BNC. The vertical plug-in contains the following controls: *volts/div, position, mode, add,* and *trigger*.

Volts/div. These controls (one for each channel) control the deflection factor for a specific channel, A through D. A vernier control is concentric to each of these controls, but the calibration is accurate only when the vernier control is turned off. If the vernier knob on any volts/div control is turned on, an *uncal* light turns on to warn the operator not to trust the deflection factor.

Position. These four controls, A through D, set the position of the four traces anywhere on the screen. They move the traces up and down. Concentric to each position control is a switch marked *off-on-inv*. These controls turn the respective channels on and off; in the INV *(invert)* mode, the control turns the waveform upside down.

Mode. Used when more than one channel is in use to determine whether sweep is to be chopped or alternate (see Sec. 8-14).

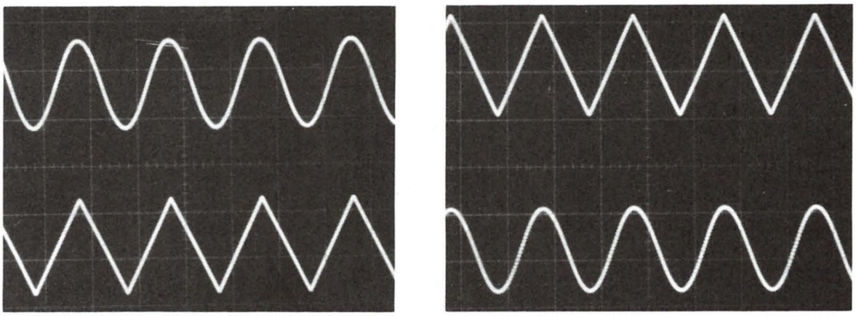
Add. These two controls add channels A and B (A + B) and C and D (C + D), respectively. When used with the INV mode, then A − B and C − D functions are created. This feature allows the inputs to be used differentially.

Trigger. These five controls select the source of the trigger signal. Any channel A through D may be used as the trigger source, regardless of whether or not that channel is displayed on the CRT. In the *comp* (composite) mode, each channel is triggered independently.

Note that the vertical position controls allow the user to place the waveform anywhere on the screen of the CRT. Figures 8-19(a) and 8-19(b) show how two waveforms, one from each channel, may be interchanged.

The horizontal time base plug-in is the unit in the lower right position on the main frame in Fig. 8-18. The controls on the time base plug-in are *time/div, trig level, reject (hf/lf), pos/neg, int/ext, ac/dc, hold off, auto/norm, single, reset,* and *delay*.

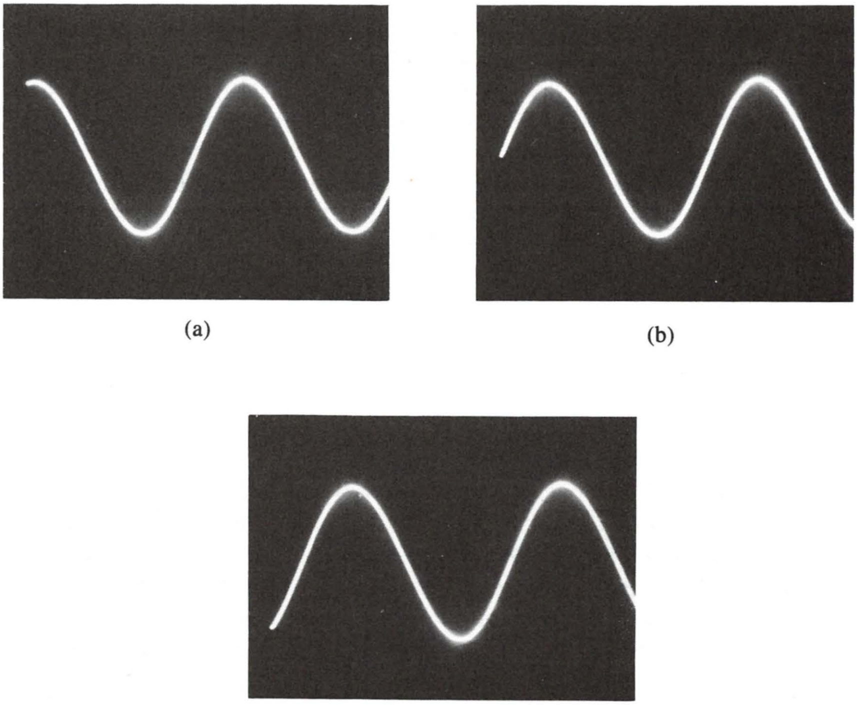
Time/div. This rotary switch selects the sweep speed. In this particular model the range is from 0.05 μs/div to 1 s/div.



(a) (b)

Figure 8-19. Position controls for each channel allow the traces to be interchanged.

Trig level. This trigger level control determines how much amplitude is required of the vertical signal before the sweep is triggered. Figure 8-20 shows the effect of this control. In Fig. 8-20(a) it is adjusted to the "+" side, so the sweep is triggered at the positive peak of the input signal. In Fig. 8-20(b) the



(a) (b)

(c)

Figure 8-20. Varying the trigger control changes position on input waveform where sweep begins.

control is set to "0," so the trigger occurs near the zero-crossing of the vertical input signal. In Fig. 8-20(c) the control is set to the "−" side, so the triggering occurs on the negative peak of the input signal.

Reject (hf/lf). These controls reject signals that could interfere with proper triggering on desired signals. The *hf reject* attenuates signals above 15 kHz, while the *lf reject* attenuates signals below 15 kHz.

Pos/neg. Selects the positive or negative slope of the input signal for triggering.

Int/ext. Selects internal (i.e., vertical amplifiers) or external trigger signals.

AC/DC. In the ac position of this switch signals below 20 Hz are attenuated, and dc is blocked. Only ac signals will trigger the sweep. This feature allows you to trigger on ac signals that are riding on dc levels.

Hold off. Adjusts the time delay between sweeps. The *hold off* control will prevent, i.e., "hold off," triggering for at least one entire sweep on all ranges of the time/div switch.

Auto/norm. In the *auto* position the sweep will trigger even if no input signal is present, but will trigger properly if an input signal is present. In the *norm* (normal) position the sweep will not trigger unless an input signal is present.

Single. When the auto/norm switch is in the *auto* position, and the *single* button is in, then pressing the *reset* button causes a single sweep of the CRT. In the *norm* position of the auto/norm switch, with the *single* button in, then one sweep occurs when a normal triggering signal is present. In this mode the *reset* button arms the trigger and lights an indicator lamp.

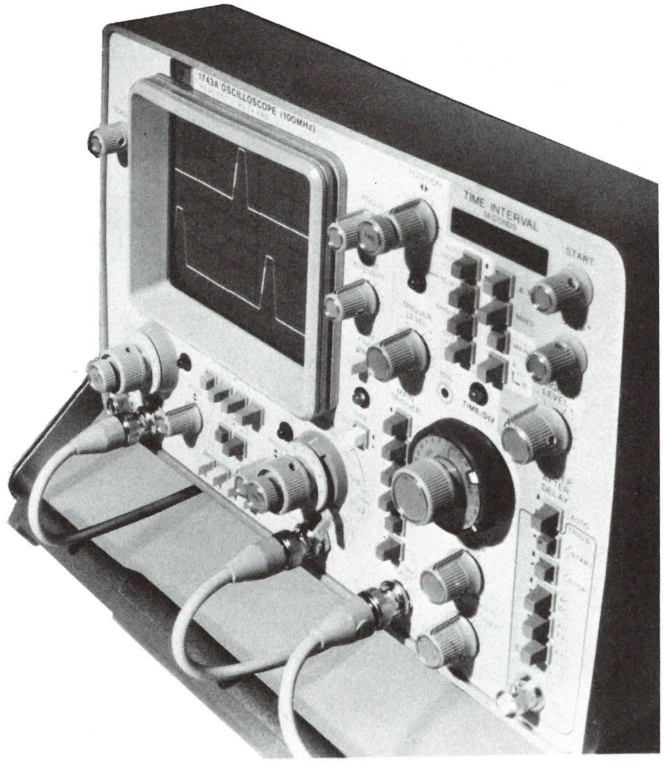
Reset. See discussion above on *single* button.

Delay. Sets the time delay between triggering and the actual beginning of the sweep. This control is especially useful for viewing small sections of a waveform, while triggering on the main signal.

8-19 DELTA-TIME OSCILLOSCOPES

Some oscilloscopes are equipped with a Δ-time feature that allows the measurement of the time interval between two points on the applied waveform. The model shown in Fig. 8-21(a) uses an internal electronic period counter to measure time intervals. The numerical readout is in a small window on the oscilloscope front panel. Markers on the CRT screen, in the form of enhanced trace brightness, allow the user to select the two points on the input waveform between which the measurement is made.

In the second example, shown in Fig. 8-21(b), the measurement is made by a digital voltmeter mounted on top of the unit. A voltage proportional to the interval is available on the back panel.



(a)

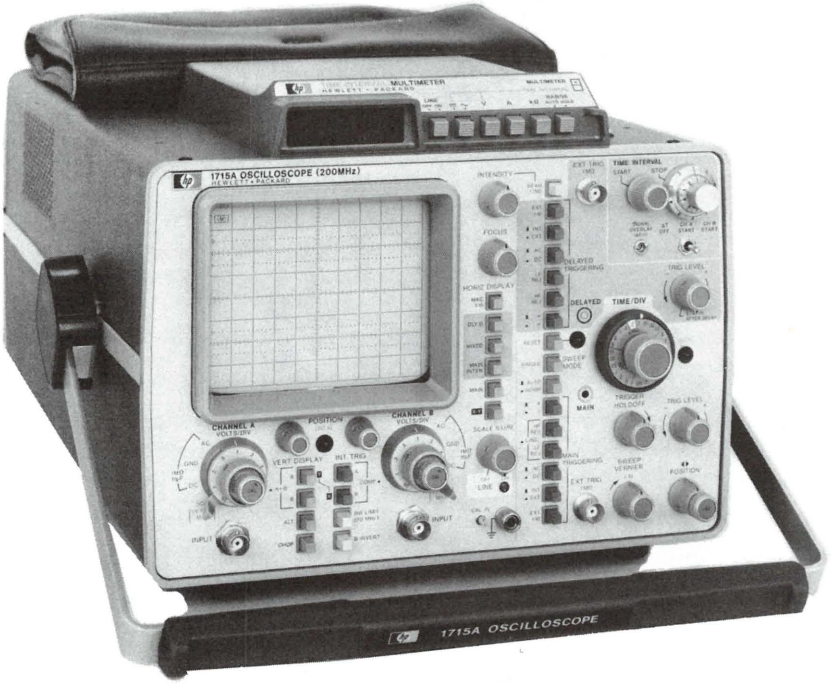
Figure 8-21. (a) Digital oscilloscope. (Courtesy of Hewlett-Packard.)

8-20 OSCILLOSCOPE CAMERAS

The oscilloscope is capable of measuring very rapid repetitive or transient waveforms. But the waveform is viewed only temporarily. Repetitive waveforms exist only as long as all of the instruments and circuit that produce them are operating, whereas transient waveforms exist only for a fleeting moment unless a storage oscilloscope is used. But even storage images fade in an hour or so.

Also, in the oscilloscope there is no provision for making a hard copy—that is to say, permanent records of the waveforms. Such records document the performance of circuits, so are deemed very important.

The solution to this problem is to make a photograph of the waveform on the oscilloscope screen. Although a few life sciences researchers and clinicians in medicine use a special 35-mm camera, equipped with a close-up lens, 100-foot magazine, and motor drive, the vast majority of oscilloscope photography is done on Polaroid® film, which gives the advantage of an instant print.



(b)

Figure 8-21. ***(cont.)*** (b) Add-on electronic voltmeter creates a digital oscilloscope. (Courtesy of Hewlett-Packard.)

There are a number of different camera options available. The Polaroid Corporation, for example, offers their Model CR-9 hand-held camera. Both Tektronix and Hewlett-Packard manufacture various models that mount directly on the oscilloscope (see Fig. 8-22). Some of these cameras are manually operated, and others are electrically operated. The model shown in Fig. 8-22 includes a viewing hood so that the operator can see the waveform on the CRT screen before he takes the picture.

Although special oscilloscope films are used, most cameras take the standard Polaroid film pack, so will accept any black and white film that has an ASA rating of 3000 (e.g., Polaroid types 107, 085, etc.).

An important parameter in considering any oscilloscope photographic system is the *writing speed,* which is usually expressed in centimeters per microsecond. This parameter measures the effectiveness of the entire system, and is determined by factors such as the camera lens, CRT phosphor, beam spot size, intensity, etc.

Postfogging of the film can result in as much as 2X or 3X increase in writing speed. This technique exposes the film to a light source after it has been exposed to the CRT trace. On some oscilloscopes this is accomplished by turning

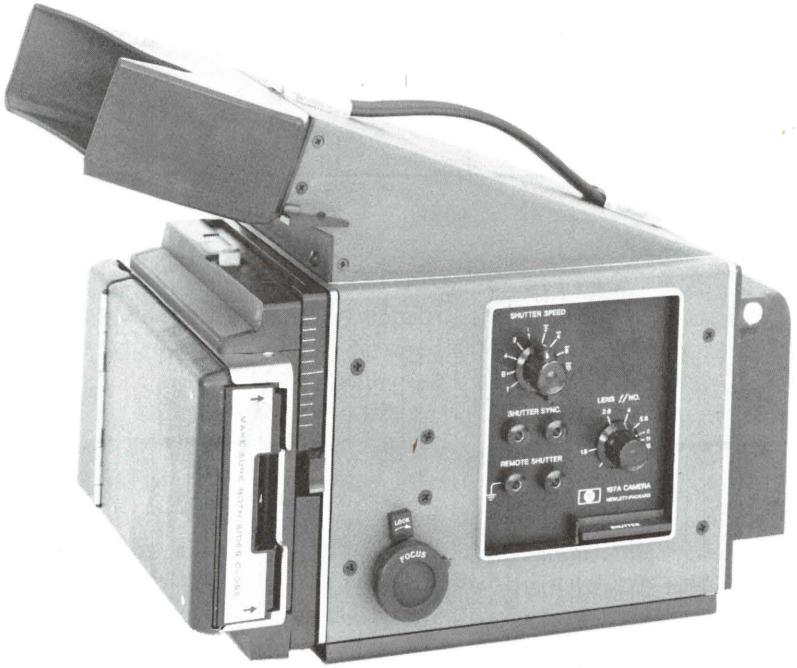


Figure 8-22. Oscilloscope camera. (Courtesy of Hewlett-Packard.)

on the CRT storage flood gun, whereas in others the graticule illumination is used. In the H-P model 197A shown in Fig. 8-22, an ultraviolet light source inside the camera is used.

8-21 SAMPLING OSCILLOSCOPES

The vertical amplifier bandwidths of even the costliest oscilloscopes extend to only 250 mHz, or so. Similarly, sweep speeds only go up to possibly 0.1 μs/div. When one is attempting to view pulse waveforms in the over-500-mHz region, it is necessary to use a *sampling oscilloscope*.

A simple view of sampling is given in Fig. 8-23(a). The input signal is too fast, i.e., its frequency is too high, to be displayed directly on the CRT. But if the waveform is repetitive, then we may take amplitude samples from different pulses in the wave train, and display them on a CRT screen. After several sweeps, the storage oscilloscope will have averaged a sufficient number of samples that a good representation of the waveform will appear on the CRT screen. If the sampling rate is within the range that the oscilloscope can handle, then the sampling effectively extends the range of operation well into the gigahertz region.

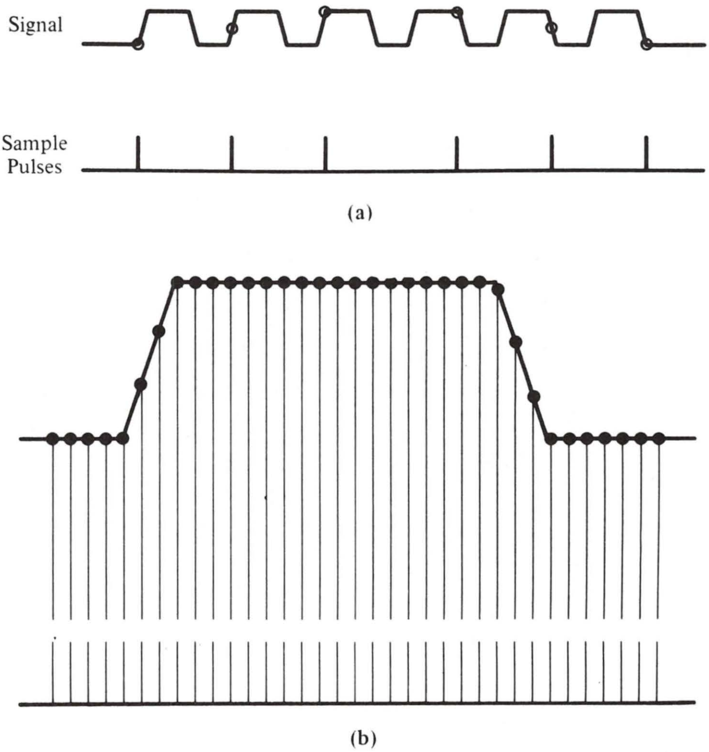


Figure 8-23. (a) How a sampling oscilloscope operates. (b) Reconstructed waveform.

8-22 Z-AXIS MODULATION

The intensity, or brightness, of the CRT trace is dependent upon the strength of the electron beam, and the kinetic energy imparted to each electron in the beam by the accelerator electrodes. We can vary the strength of the electron beam relatively easily, so by this method we obtain control over intensity.

The X-axis of the CRT is the horizontal channel, and the Y-axis is the vertical channel. It is common practice to refer to the intensity modulation channel as the "Z-axis." Many oscilloscope models contain circuitry to vary the intensity through an external jack that is usually labeled *Z-axis, Z-input,* or *intensity modulation*.

Figure 8-24 shows two popular methods for accomplishing Z-axis, or intensity, modulation. In Fig. 8-24(a) the CRT cathode is held constant by a bias supply, and the Z-axis signal is applied to a control grid. In Fig. 8-24(b), on the other hand, we have exactly the opposite situation; the control grid bias is constant, and the Z-axis signal is applied to the cathode.

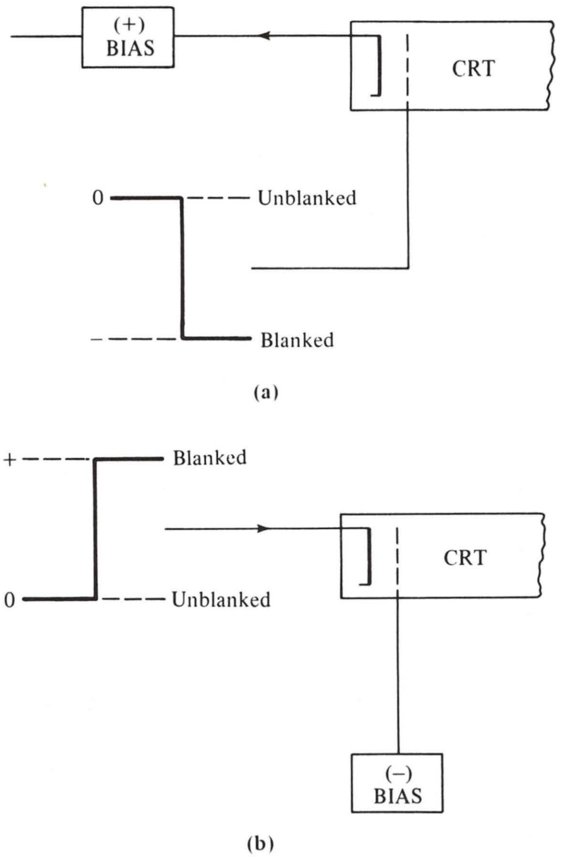


Figure 8-24. (a) Grid blanking. (b) Cathode blanking.

In most oscilloscopes using a Z-axis input, it is merely necessary to reduce the intensity (using the normal front panel *intensity* control) so that the beam is just barely extinguished. If this is done when the Z-axis signal is zero, then applying a signal will brighten the CRT trace.

8-23 SUMMARY

1. The CRT oscilloscope allows evaluation of a signal's waveshape, measurement of its frequency, period, duration, phase, and other factors.
2. Two types of deflection are used: magnetic and electrostatic. Most laboratory and service oscilloscopes use the electrostatic because it is capable of higher frequency operation.

3. In a Y-T oscilloscope the horizontal axis is connected to a time base (sawtooth) signal, and the vertical input (i.e., Y-axis) is driven by the input signal.
4. The vertical section is calibrated in volts/division, while the horizontal is calibrated in time/division.
5. A storage oscilloscope will retain the trace for up to an hour, or until it is intentionally erased.

8-24 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

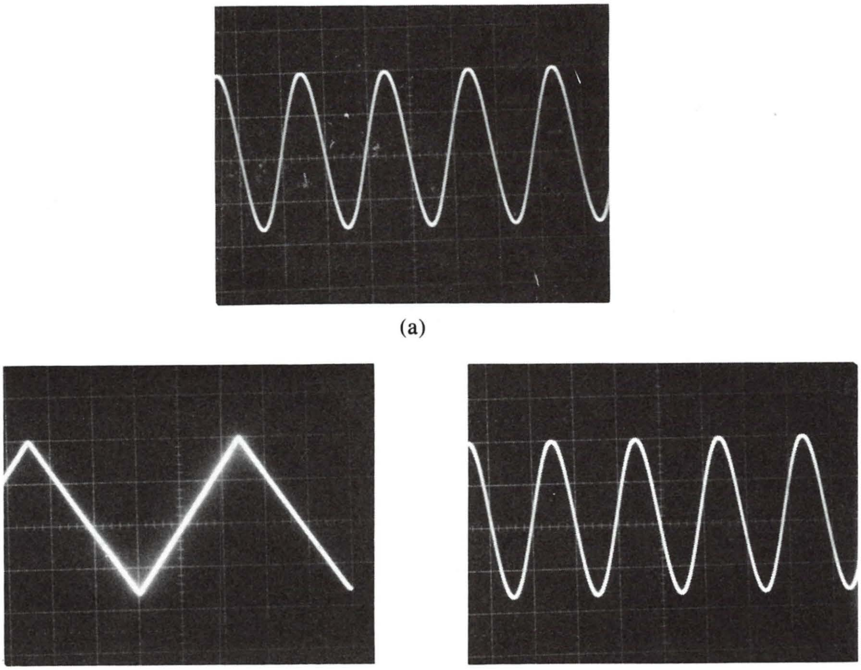
8-25 QUESTIONS

1. In addition to waveshape, it is also possible to measure the following parameters of a signal displayed on an oscilloscope: ____________, ____________, ____________, and duration.
2. Name the principal components of an electrostatic cathode ray tube.
3. List two different types of CRT, classified by the deflection method used.
4. Describe the construction of an electron gun.
5. What purpose is served by the acceleration electrode in a CRT?
6. Light on the CRT screen is created by ____________ bombardment of a ____________ surface.
7. Which two elements in a CRT are responsible for *focusing* the beam?
8. Many oscilloscopes have a positively charged electrode located between the deflection plates and the screen. This ____________ ____________ electrode is kept at a potential of several kilovolts.
9. The electrode in Question 8 is known to improve ____________ .
10. An X-Y oscilloscope can be used to produce ____________ patterns that determine the relative frequency between the vertical and horizontal frequencies.
11. In a Y-T oscilloscope the horizontal deflection system is driven by a ____________ waveform.
12. The time base controls on a free-running oscilloscope are usually calibrated in units of ____________.
13. A ____________ blanking pulse is used in a Y-T oscilloscope to eliminate the line on the CRT during retrace.

14. A ___________ sweep oscilloscope holds off the beginning of a sweep until the input signal begins.
15. A quality oscilloscope will usually have a vertical attenuator that is calibrated in ___________ / ___________, and a time base calibrated in ___________ / ___________.
16. Describe the purpose of the delay line in the vertical amplifier of a triggered-sweep oscilloscope.
17. Describe in your own words what is meant by "delayed sweep."
18. What advantages are conferred by delayed sweep?
19. Describe in your own words the difference between *focus* and *astigmatism*.
20. Define *vertical sensitivity* and *vertical deflection factor* in your own words.
21. Define *vertical bandwidth*. List two ways this specification is sometimes abused in advertising.
22. Define the rise time of a pulse.
23. Electronic ___________ is used to create a dual-beam oscilloscope, even though the CRT has but one electron gun.
24. Define ALT and CHOP modes as they apply to dual-beam oscilloscopes.
25. Three types of storage CRT are ___________, ___________, and ___________.
26. Describe how a digital storage oscilloscope works. Use a diagram if necessary.
27. A *flood gun* is used in ___________ CRTs.
28. What is meant by "variable persistence"?
29. Describe the action of an X5 horizontal magnifier.
30. The *lf reject* button in the trigger circuit ___________ frequencies below a specified point.
31. Describe the differences between the *normal* and *auto* triggering modes.
32. Most oscilloscope photography is done with a Polaroid film with an ASA rating of ___________.
33. What factors affect the photographic writing speed? What units are generally used to measure writing speed?
34. Describe the technique of postfogging, and what it does.
35. Gigahertz range signals may be viewed on a ___________ oscilloscope.

8-26 PROBLEMS

1. In Fig. 8-25(a) find the frequency of the waveform if the time base is set to 0.5 μs/div.
2. In Fig. 8-25(a), find the rms amplitude of the waveform if the vertical attenuator is set to 2 V/div.



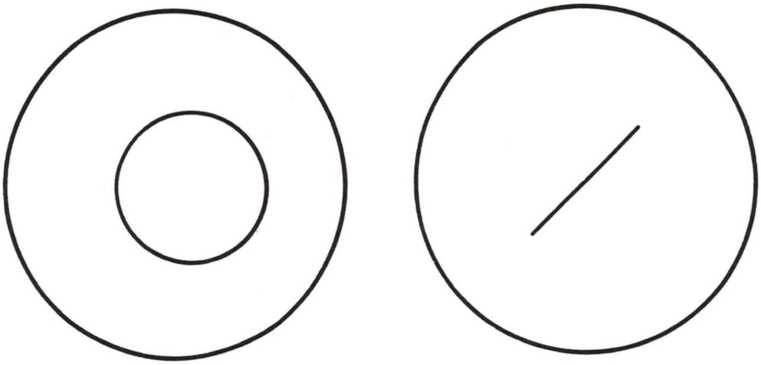
(a)

(b)

(c)

Figure 8-25.

3. In Fig. 8-25(b) find the period of the waveform if the time base is 10 ms/div.
4. In Fig. 8-25(b) find (a) the amplitude if the vertical attenuator was set to 1 V/div, and (b) the slope of the leading edge of the waveform.
5. In Fig. 8-25(c) find the frequency if the time base is set to 2 μs/div.
6. In Fig. 8-25(c) find the amplitude if the vertical attenuator is set to 5 V/div.
7. A 1-kHz sine wave exactly fits onto a 10-cm-wide CRT graticule. What is the setting of the time base knob in time/div units?
8. In Fig. 8-26(a), the Lissajous pattern shows a ____________-deg phase difference between vertical and horizontal signals.
9. What is the phase angle of the signals in Fig. 8-26(b)?
10. The Lissajous pattern in Fig. 8-27 is produced when a 41-kHz signal is applied to the vertical channel and a signal of unknown frequency is applied to the horizontal input. Find the frequency of the unknown signal.
11. Find the bandwidth in hertz required of an oscilloscope vertical amplifier if the rise time of the input pulse is 26 nanoseconds.



(a) (b)

Figure 8-26.

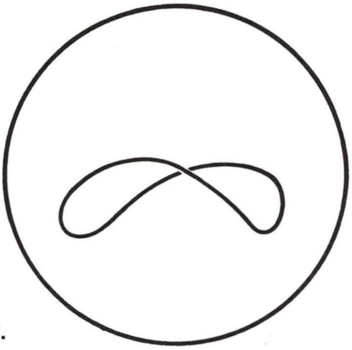


Figure 8-27.

12. Consider Fig. 8-28. A 3.58-mHz burst signal eight cycles in duration is on the "back porch" of an 11-μs pulse. If a delayed sweep oscilloscope is used to view this signal, without the main pulse, we would set the delay time to ____________. Which of the following time base speeds is most suited to viewing the burst: 1 μs/div, 2 μs/div, 5 μs/div, or 1 ms/div?

8-27 REFERENCES

1. Clyde Herrick, *Oscilloscope Handbook*. Reston, Va.: Reston Publishing Co., 1974.
2. Derek Cameron, *Advanced Oscilloscope Handbook*. Reston, Va.: Reston Publishing Co., 1977.

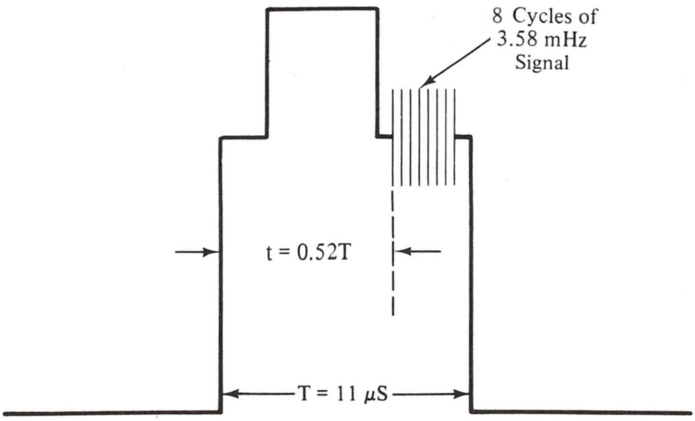


Figure 8-28.

chapter 9

SIGNAL GENERATORS

9-1 OBJECTIVES

1. To learn the basic types of signal generator.
2. To learn the types of signals available.
3. To become familiar with certain commercial models.
4. To learn appropriate applications for each type of signal generator.

9-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. Name two types of oscillator circuit commonly used in audio generators.
2. Give two different methods for creating square waves from a sine wave source.
3. What type of stage is needed to create triangle waves from square waves?

9-3 SIGNAL GENERATORS

A *signal generator* is an instrument that provides a controlled output signal for use in testing or aligning, or in measurements on other circuits or equipment. Like most electronic devices made in the past ten years or so, the modern signal generator is considerably more sophisticated than its predecessors. Yet, at the same time, some recent products are little more than solid-state versions of designs that originated prior to World War II. It seems that the effect of solid-state technology has been to update traditional designs, and to prompt even newer designs and techniques. Regardless of the techniques used in the design, however, solid-state circuits have allowed manufacturers to make packages smaller; what was once several cubic feet of cabinet is now in a box small enough to get lost on a crowded workbench.

It used to be easy to discern the different classes of signal generators. The classes were relatively rigid, with only a few exceptions, so one knew exactly what performance and capability to expect in any given product. But today, performance and specifications overlap to such an extent that classification becomes almost hopelessly difficult. There are very often several instruments in any given laboratory or shop that will do any given job. With this situation in mind, let us examine some of the traditional classes, even while caution reminds us of the many exceptions.

Audio generators. This class of instruments traditionally covers the range 20 Hz through 20 kHz, although a few models would produce signals up to 100 kHz. Audio generators always produce sine waves, and most also produce square waves. An audio signal generator typically uses either output metering, or a calibrated output attenuator, and produces controllable signal levels down to approximately –40 or –50 VU.

Function generators. These generators typically cover at least the same frequency range as audio signal generators, but most modern designs have extended frequency ranges. Some, for example, operate as low as 0.001 Hz, while others operate to 11 mHz. One is known to operate to 80 mHz. A very common frequency range is 0.01 Hz to 11 mHz.

The principal difference between a function generator and an audio generator is in the number of output waveforms that are produced. The audio signal generator produces only sine waves and square waves, while almost all function generators produce these basic waveforms plus triangle waves. Some function generators also produce sawtooth, pulse, and nonsymmetrical square waves.

The output attenuator forms the other principal difference between these two types of signal generator, although this is not a universal situation. Most audio generators use a precisely calibrated or metered output control capable of producing known output amplitudes, even at very low levels. The function generator, however, typically has a coarse, uncalibrated, output level control. Although the specifications sheet will list the output range as 0 to 10 volts rms, it is very difficult to obtain precise amplitude levels, especially at levels less than a few hundred millivolts. Both types of signal generator, however, will typically be rated to have a 600-ohm output impedance.

RF generators. An RF signal generator produces output frequencies in the "over 30 kHz" region. Most high-quality RF signal generators have a precision output attenuator that allows setting of output levels from under 1 μ V to 100,000 μ V. RF signal generators typically have a 50-ohm output impedance. Few RF signal generators cover the entire spectrum from "over 30 kHz to daylight." Some will, however, cover extremely wide ranges, with 1 mHz to 400 mHz being very easy to obtain in one instrument. Table 9-1 shows the various RF

Table 9-1 RF FREQUENCY RANGE DESIGNATIONS

Designation	*Frequency Range*
Audio*	20–20,000 Hz
Very low frequency (VLF)	20–100 kHz
Low frequency (LF)	100–500 kHz
Medium wave (MW)	500–3000 kHz
High frequency (HF)	3000–30,000 kHz
Very high frequency (VHF)	30–300 mHz
Ultra high frequencies (UHF)	300–1000 mHz
Microwaves	(over 1000 mHz)†

*Audio waves are actually *acoustical,* but since they cover the same frequency range, the word "audio" is used to describe electrical oscillations in the 20–20,000-Hz range. When an electrical oscillation, of whatever frequency, is radiated into space as an electromagnetic wave, it is *radio.* The 13-kHz Navy submarine communications stations, therefore, are producing RF signals, even though most frequency spectrum charts show 13 kHz as "audio."

†Some call frequencies over 10,000 mHz "SHF" for super high frequencies, while others call them UHF for unbelievably high frequencies.

frequency ranges and the generally accepted names for each range. Note that most signal generators will actually cover overlapping portions of several of these ranges.

RF signal generators operating in the over 1 gigahertz (i.e., 1000 mHz) region are generally called *microwave signal generators.*

9-4 AUDIO GENERATORS

The audio oscillator is used typically in tests on high fidelity, public address, and other audio equipment. These generators typically produce sine waves and symmetrical square waves at frequencies between 20 Hz and 20 kHz. Most use a 600-ohm output impedance, and produce output levels from −40 dBmW to +4 dBmW.

There are two methods of frequency selection typically used in audio signal generators, *continuous* and *step.* The continuous type of dial is much like the selector dial of a radio receiver in that you must turn a knob to the desired frequency. Many such audio generators have a scale that reads 20 to 200 (or, alternatively, 2 to 20), and a *range* selector switch determines whether the output frequencies will be 20 to 200 Hz, 200 to 2000 Hz, or 2000 to 20,000 Hz. In a step-tuned generator these controls are replaced by a rotary or pushbutton switch bank. As many as four decade switches might be used, although three is a more common number. These will be marked 0 through 100, 0 through 10, and 0.1 through 1.0 in decade steps. A *multiplier* switch determines whether the actual frequency will be ×1, ×10, ×100, or ×1000 times the frequency indicated on the selector switches. Consider, for example, a generator in which the tens switch is set to 60, the units switch to 3, and the fraction switch to 0.4, and the multiplier is set to ×100. The output frequency will be $(60 + 3 + 0.4) \times 100 = (63.4)(100) = 6340$ Hz.

9-5 AUDIO OSCILLATOR CIRCUITS

Although there are many different oscillator circuits that can be made to function in the audio range, only a few are really practical for use in audio signal generators. An LC oscillator, for example, can be made to work in the audio range, but there are certain mechanical problems that make them unusable; the LC components become excessively large. With the exception of the heterodyne technique, most audio oscillators used in audio signal generators tend to be RC designs.

An RC *phase shift oscillator* is shown in Fig. 9-1. In any oscillator, the criteria for oscillation are (1) a 360-deg (0-deg) phase shift and (2) gain greater than unity around the feedback loop. An active element such as a transistor or operational amplifier provides the needed gain. In Fig. 9-1 a JFET is used as the active element. Since $Q1$ is operated in the common-source mode, it provides a 180-deg phase shift between input and output. But to satisfy the criteria for oscillation, an additional 180 deg of phase shift at the desired frequency must be provided by the feedback path. The shift is provided by three 60-deg RC networks:

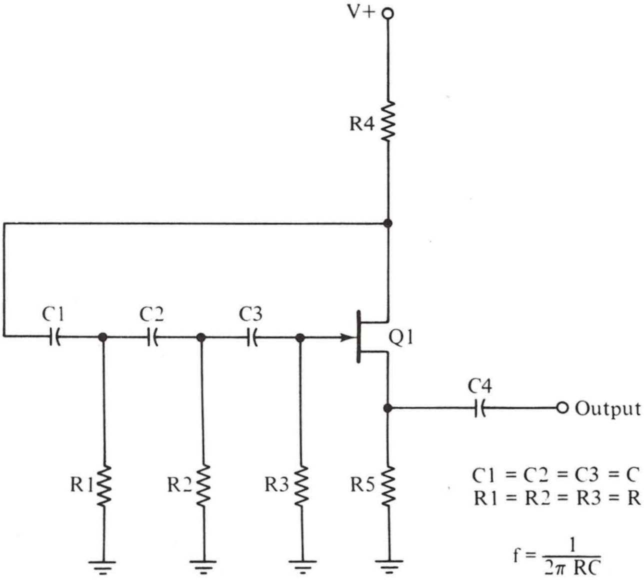


Figure 9-1. RC phase shift oscillator.

$R1C1$, $R2C2$, and $R3C3$. In some cases, $R1$ through $R3$ are potentiometers so that continuously variable frequency tuning is available. Or, alternatively, these resistors might be part of a decade resistor assembly to provide step tuning of the oscillator frequency. Capacitors $C1$ through $C3$ are part of the range switch, a different value being used for each range of frequency coverage. The oscillator's frequency of oscillation is given by

$$F_0 = \frac{1}{2\pi RC} \tag{9-1}$$

where F_0 = the operating frequency in *hertz* (Hz)
R = the resistance in ohms (Ω)
C = the capacitance in *farads* (F)

An example of a Wien bridge oscillator is shown in Fig. 9-2. A differential amplifier $A1$ is connected across the output nodes of an RC Wien bridge consisting of four legs: $R1$, $I1$, $R3C1$, and $R4C2$. At all times E_0 and $E1$ are in phase with each other, and 180 deg out of phase with $E2$. The feedback in this loop is degenerative, therefore stable, at all frequencies except F_0. Frequency F_0 is also described by Eq. (9-1). Voltage $E3$ is out of phase with E_0, but at frequency F_0,

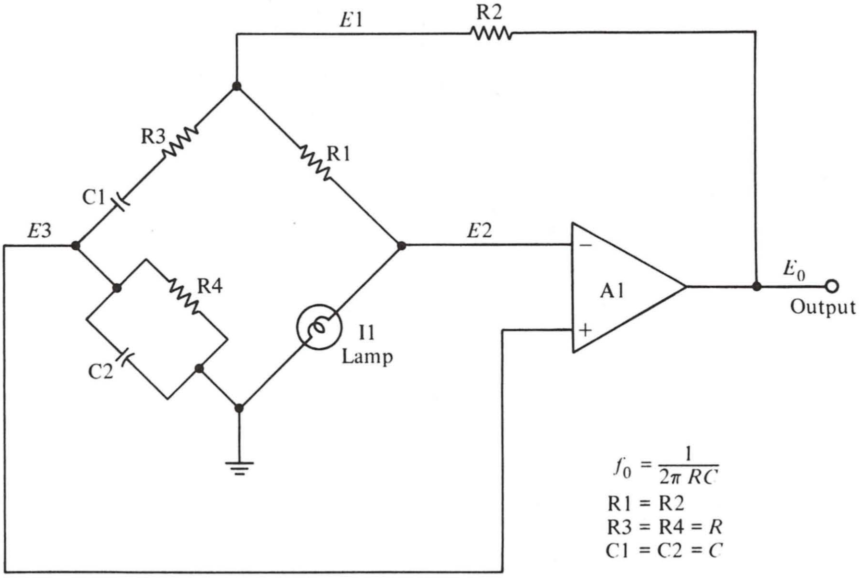


Figure 9-2. Wien bridge oscillator.

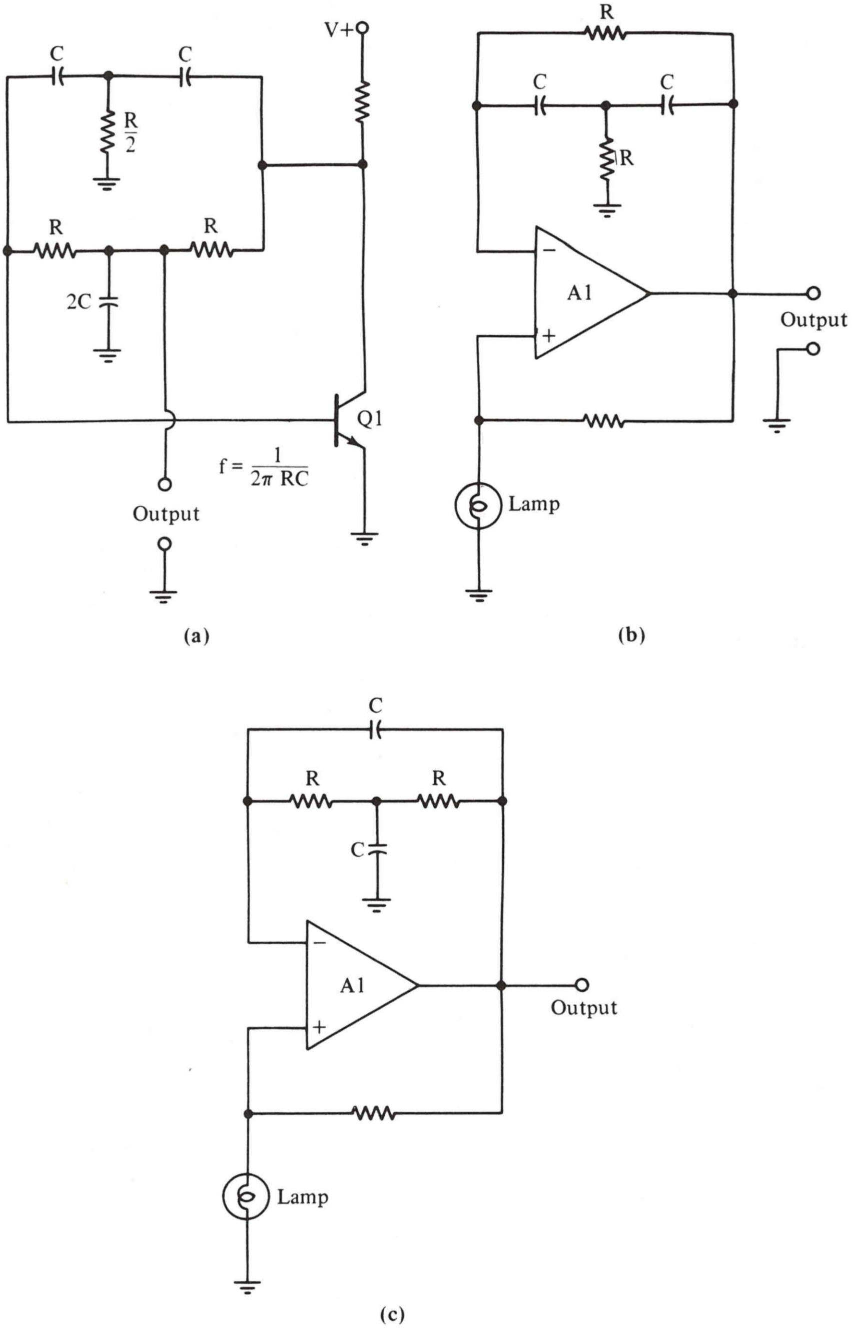


Figure 9-3. (a) Twin-T oscillator. (b) Bridged-T oscillator (resistance type). (c) Bridged-T oscillator (capacitor type).

voltages $E3$ and E_0 are in phase; i.e., $\Theta = 360$ deg. Since $E3$ is applied to the noninverting input of the amplifier, this circuit will oscillate.

Lamp $I1$ is used to stabilize the output amplitude and prevent saturation of the amplifier device. This job is accomplished by the changing resistance of the lamp as E_0 increases, causing more current to flow in the lamp filament. In most cases, $I1$ is operated at current levels below incandescence.

The Wien bridge oscillator is used in many audio signal generators because of its relatively low distortion. Although $R3/R4$ may be made variable by using either a potentiometer or decade switch and resistor bank, it is the latter that is most popular.

A twin-tee oscillator circuit is shown in Fig. 9-3(a). Transistor $Q1$ provides 180 deg of the required 360-deg phase shift, while the remaining 180 deg is provided by a twin-T *RC* network. There are two qualitative ways to describe this circuit and its operation. In one sense, we point out that there is a 180-deg phase shift through the network at frequency F_0 [Eq. (9-1)]; hence, the input of the amplifier sees an in-phase feedback signal. Some other textbooks view the circuit in quite a different way. They point out that the twin-T is a notch filter; i.e., it severely attenuates F_0, but passes all other frequencies removed from F_0. Because of this fact, we are able to note that all frequencies except F_0 are given a tremendous amount of negative feedback, so the circuit gain at those frequencies is limited. But F_0 receives very little negative feedback, because only a small amount of the signal at this frequency passes the feedback network. Of these two views, the author prefers the first.

There are two variations on the twin-T design that tend to make the circuit more stable. These circuits are collectively called *bridged-T oscillators,* and are shown in Figs. 9-3(b) and 9-3(c). The resistor-bridge is shown in Fig. 9-3(b), while the capacitor-bridge is shown in Fig. 9-3(c).

The last oscillator circuit that we will consider is the *heterodyne* design, shown in Fig. 9-4. This signal generator has fallen into general disuse in recent years, but was once very common.

In the heterodyne system, a variable frequency oscillator is mixed with a fixed frequency oscillator to produce a difference frequency $(F2 - F1)$. Most heterodyne audio signal generators used LC oscillators operating in the 50- to 500-kHz region, although a few were known that used higher-frequency oscillators. In general, though, lower-frequency LC oscillators produce more stable signals than do high-frequency oscillators. The fixed oscillator might be a fixed-frequency LC oscillator, but in most cases was a crystal oscillator.

9-6 GENERATING SQUARE WAVES FROM SINE WAVES

Square waves produced by an audio signal generator must be symmetrical, not only in the left-right sense, but across the zero volts base line as well. The square

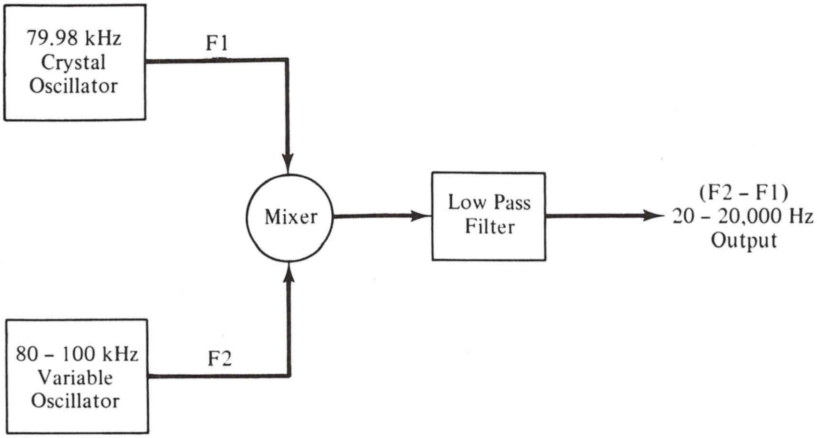


Figure 9-4. Block diagram for a heterodyne-type audio generator.

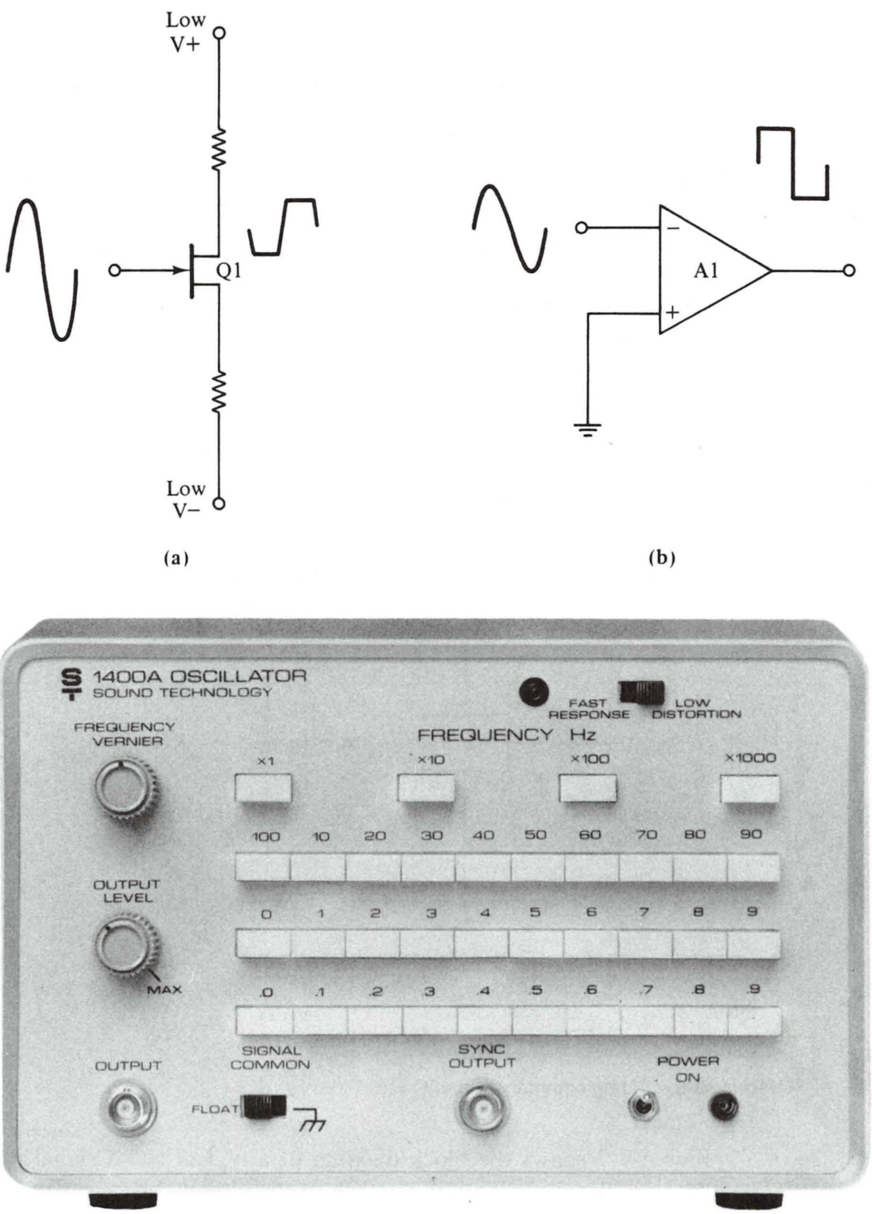
waves can be produced by an astable multivibrator that is frequency locked to the sine wave oscillator, but this is difficult to achieve reliably in practice. It is, then, considered best to derive the square waves from the sine waves directly.

Figure 9-5 shows two methods for deriving a square wave from a sine wave. In Fig. 9-5(a) we see the use of an overdriven amplifier stage, a technique that is popular in low-cost models. Transistor *Q*1 is operated with the low V− and V+ voltages, so is easily overdriven by the large-amplitude sine wave applied to the gate. The output signal is clipped severely because of the low supply voltages. Unfortunately, the rise time of the square waves produced in this manner is not as good as might be required in some applications.

A somewhat superior approach is the circuit of Fig. 9-5(b). Here we have an operational amplifier operated as a voltage comparator. The lack of feedback causes the amplifier gain to be very high (see Chapter 12). Since the noninverting input is grounded, i.e., at a potential of zero volts, the output saturates any time the input potential is more than a few millivolts greater than zero. On negative excursions of the input sine wave, the amplifier output snaps positive, whereas on positive excursions of the input signal, the output snaps negative. The bipolar power supply accounts for the base-line symmetry of the output waveform.

A high-frequency comparator will produce output square waves with very good rise times. But even this circuit can be improved by the addition of a small amount of positive feedback to speed up the output transition time, thereby improving rise time.

An example of a commercial audio signal generator featuring very low distortion output waveforms is shown in Fig. 9-5(c). This model is the Sound Technology Model 1400A, and is used primarily in testing high fidelity equipment.



(a) (b)

(c)

Figure 9-5. (a) Overdriven amplifier used as a clipper. (b) Use of a comparator as a square wave generator. (c) High-grade audio generator. (Courtesy of Sound Technology, Inc.)

9-7 OUTPUT CIRCUITS

The audio signal generator must be able to deliver as much as 10 volts rms into a 600-ohm load. It must also be capable of reducing the output level to a few millivolts, without changing the generator output impedance. Additionally, the output signal will be symmetrical about zero, so the output range will be ±10 volts rms, or ±14 volts peak. The output stage of such a signal generator will usually consist of a power amplifier operated from bipolar power supplies (see Fig. 9-6). This amplifier will be required to deliver several hundred milliwatts at very low total harmonic distortion. In any event, a buffer stage between the oscillator and the output load is always desirable, so that the operation of the oscillator is not affected by changes in the load external to the signal generator.

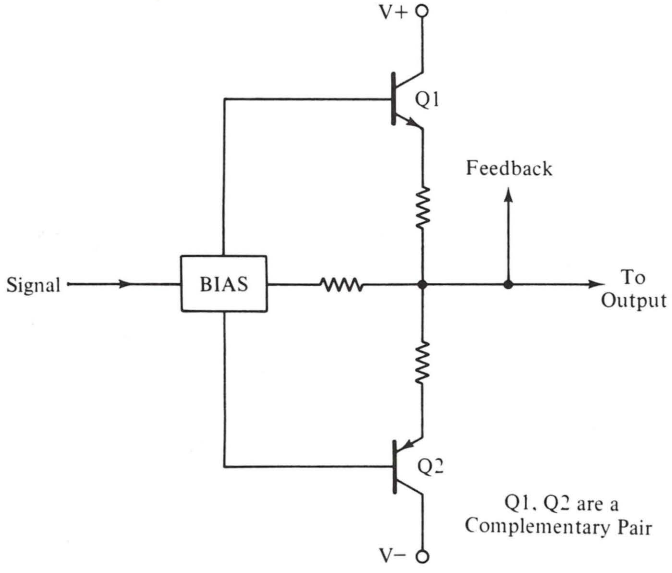


Figure 9-6. Complementary output stage.

Figure 9-7 shows the block diagram of a typical audio signal generator. The audio oscillator section will use one of the circuits given earlier. A power amplifier stage provides buffering between the load and the oscillator, and develops the output signal amplitude.

The ac voltmeter at the output is strictly optional, but in some models it is used with a *level control* to set precisely the input signal to the attenuator. Not all quality audio signal generators use this feature, so the lack of an ac output meter is not, in itself, indicative of quality.

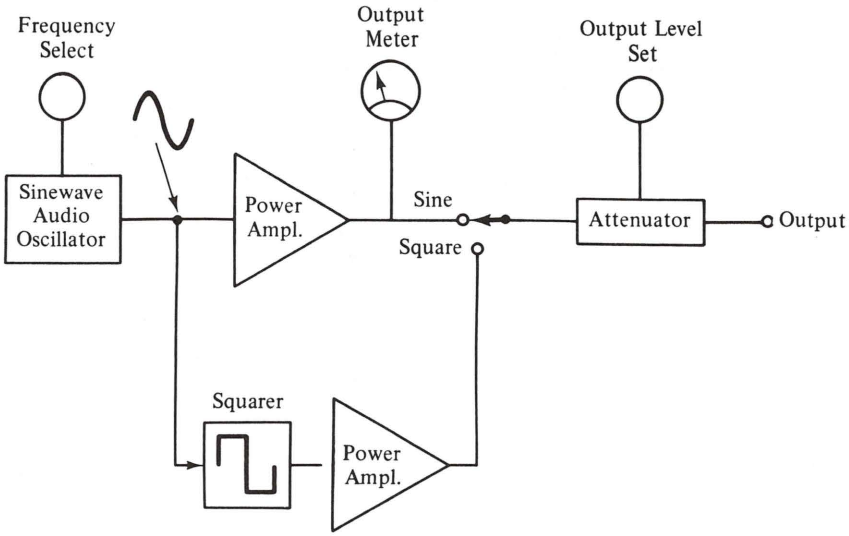


Figure 9-7. Block diagram of an audio signal generator.

Typical attenuator circuits are shown in Fig. 9-8. The circuit of Fig. 9-8(a) is used in lower grade instruments, and may take the form of a potentiometer, or switched resistor bank (as shown). This circuit is not often used in high-quality audio signal generators, because the output impedance of the instrument is not constant at all output level settings. In addition, whatever load is connected to the output affects the amplitude of the output signal severely, and each position is affected differently. This problem will lead to significant output voltage errors if nonstandard loads are used.

An unbalanced ladder network attenuator is shown in Fig. 9-8(b). This network and several popular variations are used in most commercial audio signal generators. It provides a constant output impedance, so therefore eliminates also the output voltage error of the previous design.

A square wave is created from the sine wave in a squarer circuit. In some models the square wave section has its own output terminals, while in others a *function selector* switch selects the output waveform. The output calibration is for the sine wave function and has little meaning for square waves. The square wave amplitude is usually measured in peak-to-peak volts.

9-8 FUNCTION GENERATORS

We have already mentioned that function generators and audio generators have a lot in common, so in this section we will consider only those points in which the function generator is different from the audio generator.

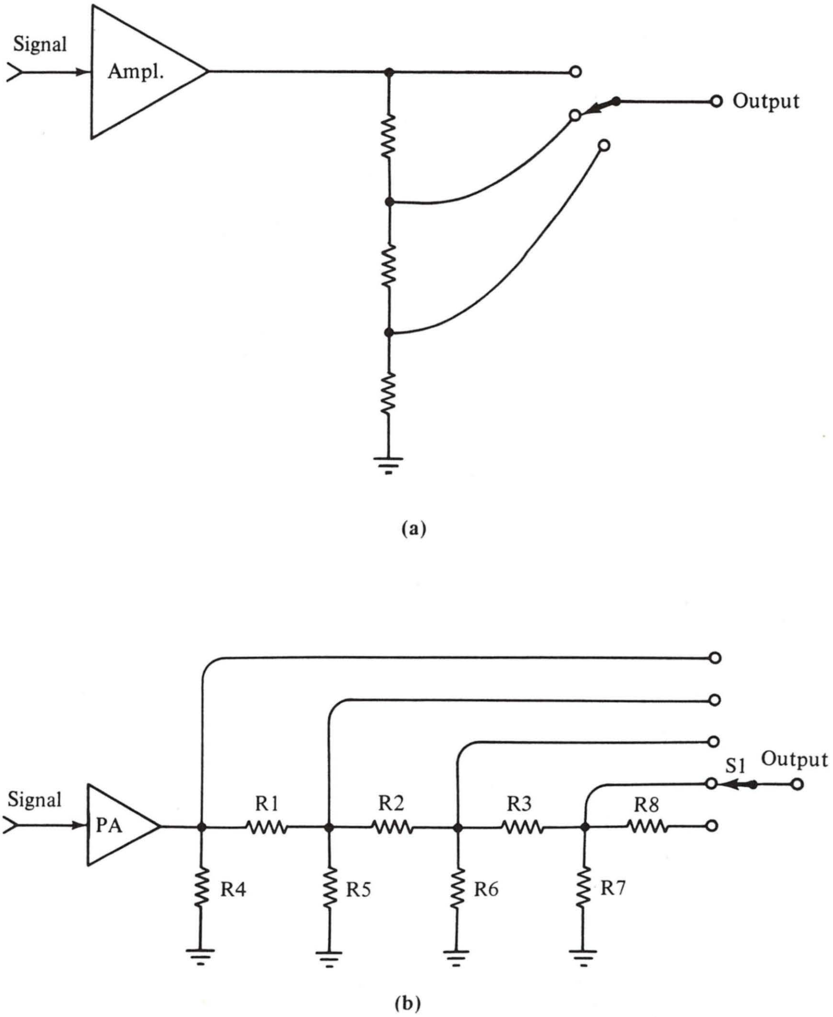
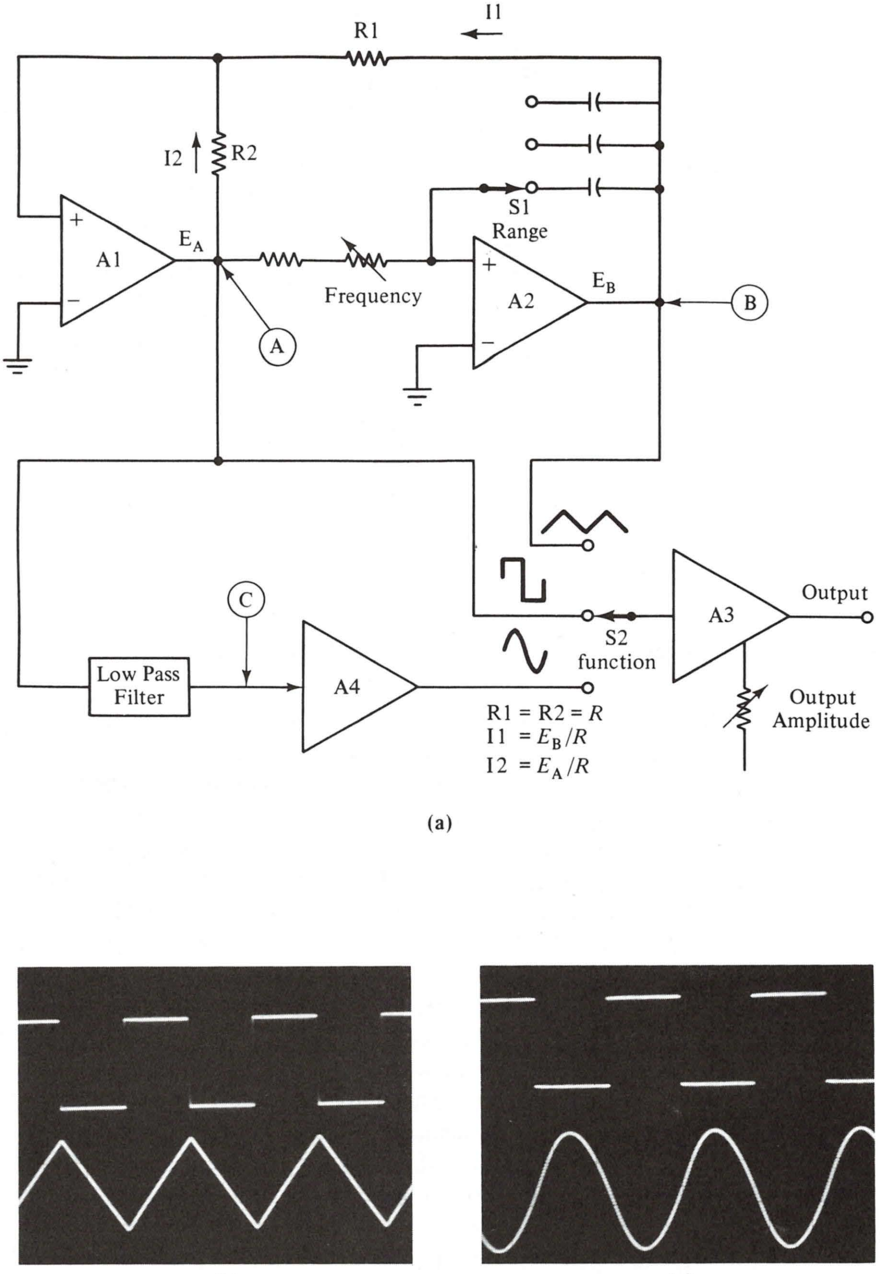


Figure 9-8. (a) Simple attenuator. (b) Constant impedance output attenuator.

Figure 9-9(a) shows a simplified schematic of a typical sine-square-triangle function generator. In most of these circuits a square wave multivibrator originates the signal. Most of the sine wave oscillators discussed in the previous section have problems with amplitude and frequency stability unless properly compensated by the designer. But certain types of square wave generator circuit have significantly better performance in this respect, and use simpler (hence less costly) circuits.

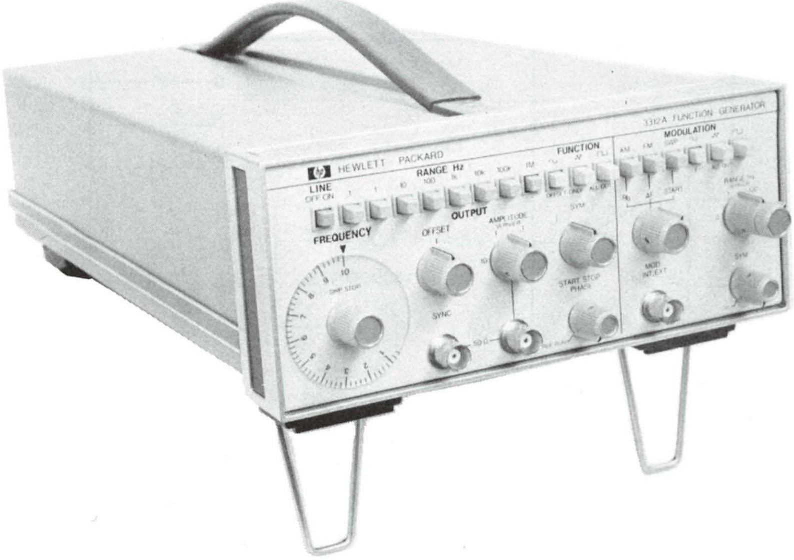


(a)

(b)

(c)

Figure 9-9. (a) Block diagram of a function generator. (b) Waveforms at points A and B in (a). (c) Waveforms at points A and C in (a).



(d)

Figure 9-9. ***(cont.)*** (d) Typical function generator. (Courtesy of Hewlett-Packard.)

In Fig. 9-9(a) amplifier $A1$ is used as a voltage comparator, while amplifier $A2$ is used as an integrator. When the circuit is first turned on, the output of amplifier $A1$ (i.e., point A) will snap high, creating current $I2$. The voltage at point A will also cause the feedback capacitor in the integrator to begin charging, thereby causing a voltage at point B to begin rising. Voltage E_B creates current $I1$. When E_B rises to the point where $I1$ is equal to $I2$, then E_A snaps back to zero, making $I2$ also zero. But E_B is still at the same level, so now $I1$ is greater than $I2$. This makes the amplifier output pass through zero to become high in the negative direction. With E_A now negative, the integrator capacitors will begin to discharge, so E_B reduces toward zero. The waveforms at points A and B are shown in Fig. 9-9(b). The upper trace shows the waveform at point A (the square wave), while the lower trace shows the waveform at point B (the triangle).

A sine wave can also be generated from a square wave, because the square wave is a composite of a fundamental sine wave and an infinite number of odd harmonics. A low-pass filter is shown in Fig. 9-9(a); it removes the harmonics, leaving only the sine wave fundamental. Figure 9-9(c) shows the result of using a low-pass filter on a square wave at point A. Although it is not apparent from the waveforms in Fig. 9-9(c), the filtering causes a substantial loss of amplitude, so an amplifier stage [e.g., $A4$ in Fig. 9-9(a)] must follow the filter section. An example of a commercial function generator is shown in Fig. 9-9(d). This instrument is the Hewlett-Packard Model 3312A.

9-9 SAWTOOTH GENERATORS

A sawtooth waveform (Fig. 9-10) consists of a ramp leading edge, followed by a rapid drop back to zero on the trailing edge. The sawtooth is used to drive the horizontal plates of an oscilloscope, and the VCO in a sweep generator.

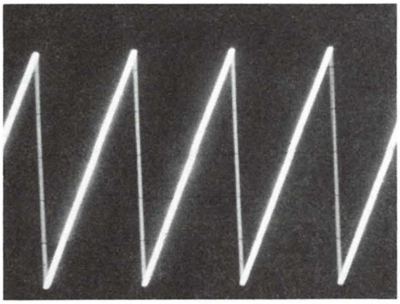
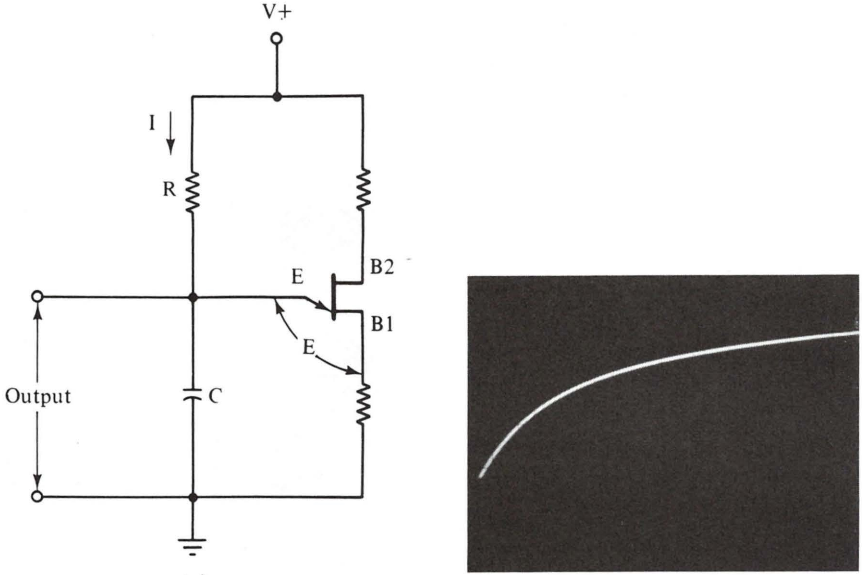


Figure 9-10. Sawtooth waveform.

Figure 9-11(a) shows one traditional sawtooth circuit, the *unijunction transistor* (UJT) relaxation oscillator. The UJT remains essentially turned off until voltage E exceeds a threshold point, at which time the UJT breaks down the base-1/emitter junction. Current I charges capacitor C until the voltage across C exceeds the UJT breakover voltage. The low resistance across the $B1$-E junction rapidly discharges the capacitor.

Unfortunately, the simple circuit of Fig. 9-11(a) produces a poor sawtooth waveform. The leading edge of a sawtooth must be a linear ramp, not an exponentially rising capacitor-charging waveform such as that shown in Fig. 9-11(b). We could linearize this waveform by using only the first 10 or 15 percent of the RC time constant, but this is difficult in UJT circuits. Another solution to the problem is to use a *constant current source* to charge the capacitor. A CCS can be a special diode-connected JFET configuration, or it may be a simple bipolar transistor circuit. The CCS keeps the capacitor charge current I constant, so the output waveform will be more linear [Fig. 9-12(b)].

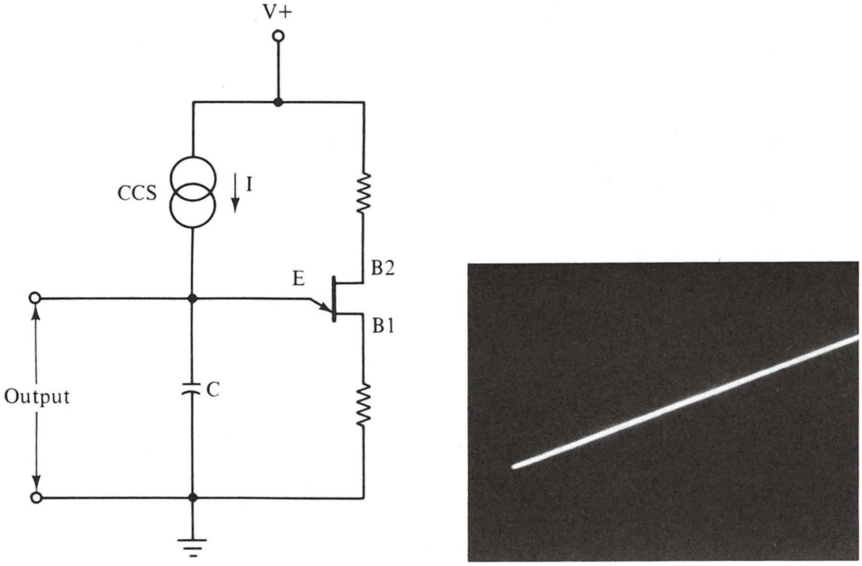
A circuit in which it is practical to use the first 10 to 15 percent of the capacitor charge curve is shown in Fig. 9-13. This circuit finds use in several modern function generators. It uses the standard Miller integrator that is used to create triangle waves, but in this case a CMOS electronic switch is connected across the feedback capacitor *(C)*. A voltage comparator is used to turn the switch on and off. When E_0 rises to a point equal to $E1$, then the comparator output snaps high, turning on the switch, which rapidly discharges C. When E_0 drops back to zero, the comparator opens the switch, allowing the procedure to repeat itself. The input to the integrator is a constant reference voltage, and this keeps both $I1$ and $I2$ also constant.



(a)

(b)

Figure 9-11. (a) UJT semi-sawtooth oscillator. (b) Waveform.



(a)

(b)

Figure 9-12. (a) Improved version of UJT circuit. (b) Waveform.

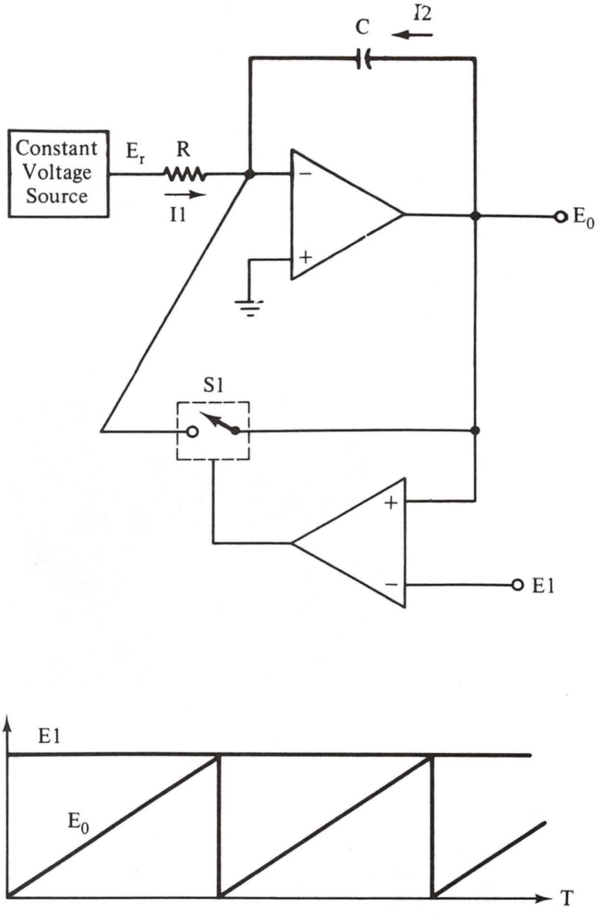
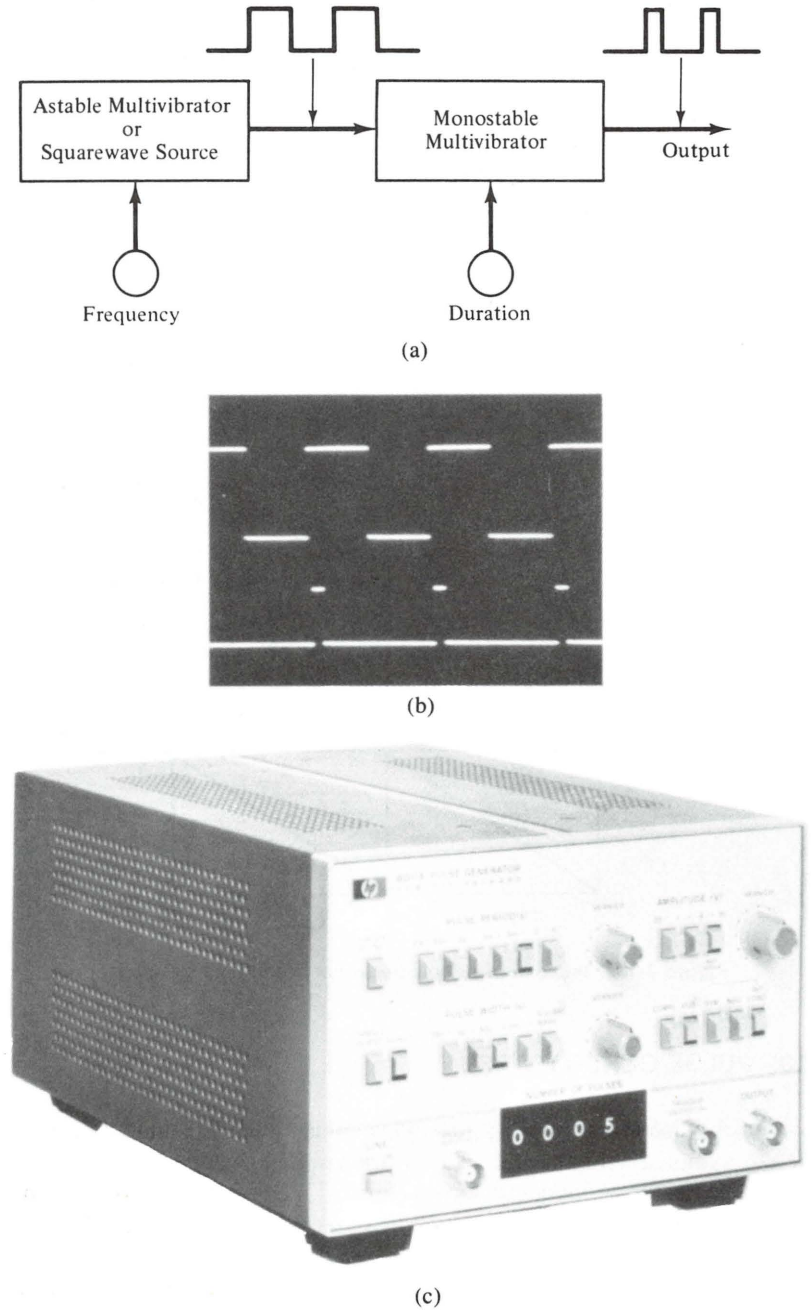


Figure 9-13. Improved sawtooth generator based on a Miller integrator.

9-10 PULSE GENERATORS

A pulse differs from a square wave in that it needs neither base line, nor left-right symmetry. Figure 9-14(a) shows one popular approach to constructing a pulse generator. A monostable multivibrator, i.e. one-shot, follows a square wave oscillator. The pulse repetition rate is set by the square wave frequency. The one-shot triggers on the leading edge of the square wave, and produces one output pulse for each input cycle. The duration of each output pulse is set by the on time of the one-shot, and may be either very short or may approach the period of the square wave. The relationship between the waveforms is shown in Fig. 9-14(b).



(a)

(b)

(c)

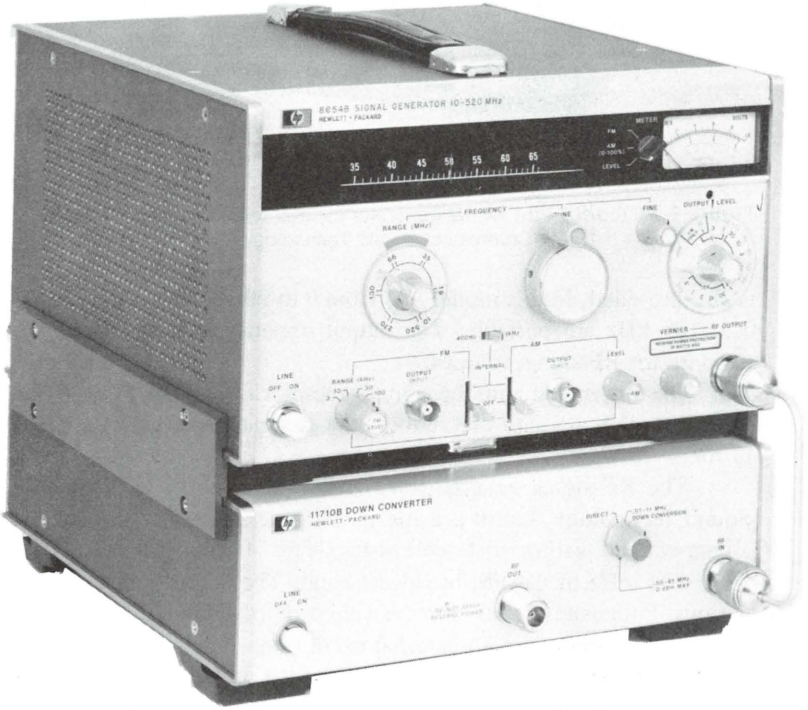
Figure 9-14. (a) Block diagram of a pulse generator. (b) Waveforms. (c) Typical pulse generator. (Courtesy of Hewlett-Packard.)

An example of a commercial pulse generator is the H-P Model 8011A shown in Fig. 9-14(c). This instrument not only allows the user to select the repetition rate, duration, and amplitude, but also the number of pulses to be output in a given burst. The setting shown in the illustration shows that a burst of five pulses will follow a trigger pulse.

9-11 RF GENERATORS

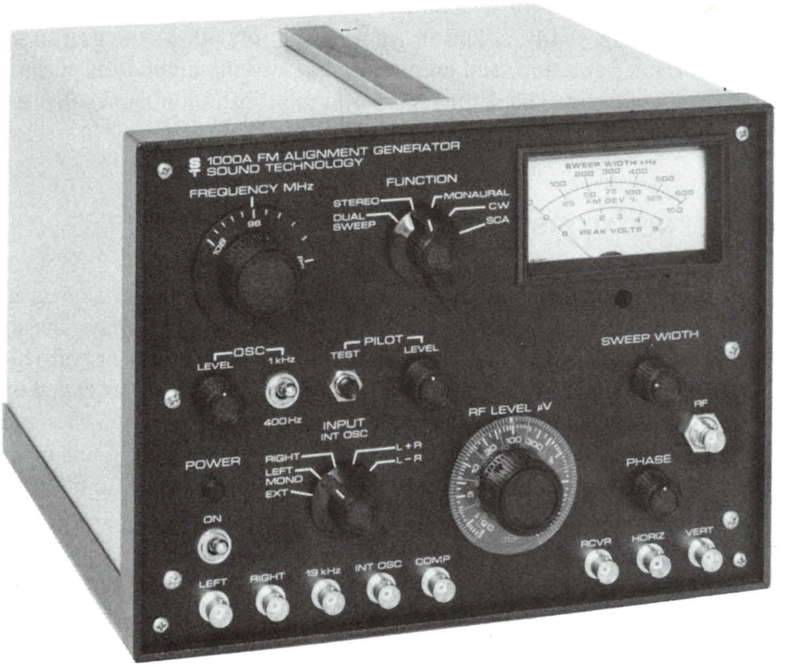
An RF signal generator uses an oscillator to produce ac signals in the region above 30 kHz. Most RF signal generators use LC oscillators as the signal source. Most RF signal generators also produce either AM or FM (or both) modulation of the output signal so that receivers and other circuits may be tested or aligned.

Figure 9-15(a) shows the Hewlett-Packard Model 8654B signal generator. It has a frequency range of 10 to 520 mHz. The output attenuator provides calibrated levels from -137 dBm to $+10$ dBm (i.e., 0.07 μV to 0.7 V). Modulation



(a)

Figure 9-15. (a) Typical high frequency to VHF RF signal generator. (Courtesy of Hewlett-Packard.)



(b)

Figure 9-15. ***(cont.)*** (b) Signal generator for alignment of stereo FM broadcast receivers. (Courtesy of Sound Technology, Inc.)

is also provided. In this model AM from 0 to 90 percent, and FM deviation from 0 to 100 kHz are possible. The output impedance is the 50 ohms considered standard for RF signal sources.

The unit mated with the signal generator is a model 11710B *down converter*. It is an optional unit that reduces the generator output to the range 10 kHz to 11 mHz.

The RF signal generator shown in Fig. 9-15(b) is the Model 1000A by Sound Technology, Inc. It is a special-purpose instrument designed for use in the alignment and testing of stereo FM receivers. The output frequency, then, is the 88 to 108 mHz of the FM broadcast band. The Model 1000A also provides the various composite signals for the stereo demodulator in the receiver, and the connections required to an external oscilloscope during alignment (Chapter 21). The Sound Technology 1000A was one of the first high-quality signal generators designed expressly for the consumer electronic service industry. Most previously popular instruments were of poor quality. A good signal generator uses double shielding; i.e., the oscillator is housed inside a shielded metal box inside of the shielded cabinet. But low-cost models often depend solely upon the cabinet for

Figure 9-15. *(cont.)* (c) Modern programmable signal generator. (Courtesy of Hewlett-Packard)

shielding. In these instruments the leakage of RF signal around the cabinet flanges is often greater than the signal coming out through the attenuator and output jack! This will easily make any measurements or alignment jobs performed suspect as to quality.

The signal source in the common RF signal generator might be any of several popular LC oscillator circuits. Both Colpitts and Hartley designs are used extensively.

The H-P 8165A signal generator is a modern instrument (see Fig. 9-15(c)). This instrument features digital readout.

The most basic RF signal generator configuration is shown in Fig. 9-16. Here we have a band-switched, tunable oscillator and buffer driving the attenuator directly. The output meter determines the level applied to the attenuator. In most cases, the meter is not calibrated (the attenuator *is*) in units, but has a calibration point, often a red line on the scale. A level control varies the gain of

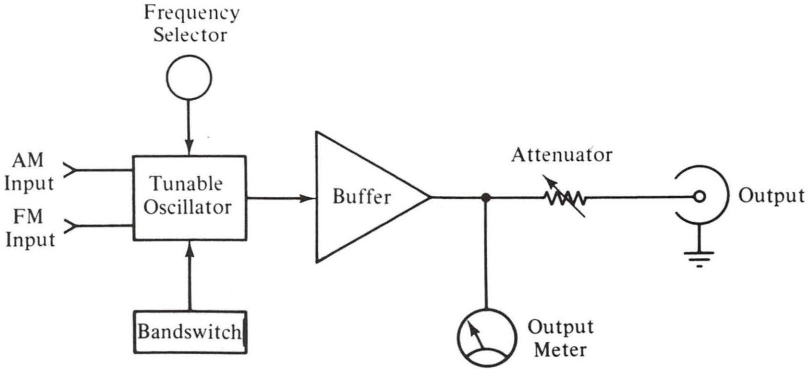


Figure 9-16. Block diagram of simple RF signal generator.

the buffer, or the signal to the attenuator, to the red line. When the output is set to the red line, then the level at the attenuator input is known, so its calibrations are valid. In some older signal generators the output frequency changes slightly when the attenuator is adjusted. But the calibrations will be accurate if the level control is then adjusted to red line.

An automatic approach to signal generator design is taken in the circuit of Fig. 9-17. The previous design required a level control to keep the output signal amplitude constant. But in Fig. 9-17 an *automatic gain control* (AGC) circuit is used. An amplitude modulator is placed between the output of the oscillator and the input of the attenuator. The oscillator signal is the "carrier" and the output of an AGC rectifier is the modulating signal. The rectifier output is proportional to the RF signal at the modulator output, so by feedback action it keeps this level constant. Since the stage is an amplitude modulator, we may also apply an ac signal (usually 400 or 1000 Hz) to this stage to provide the AM modulation function.

9-12 FREQUENCY SYNTHESIZERS

In the past decade or so a new type of RF signal generator has become popular: the *frequency synthesizer*. Various frequency synthesis schemes have been tried over the years, but all except one have suffered from various faults or design

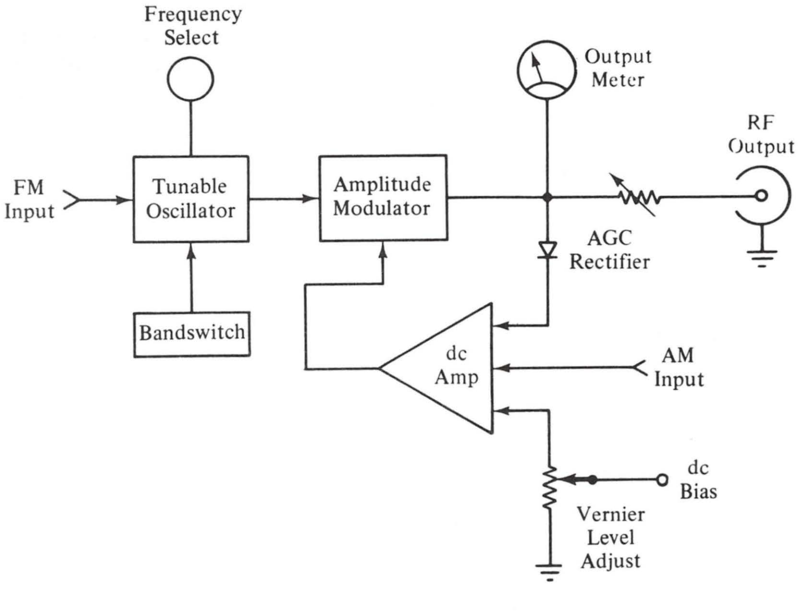


Figure 9-17. Block diagram of RF signal generator with automatic output level control.

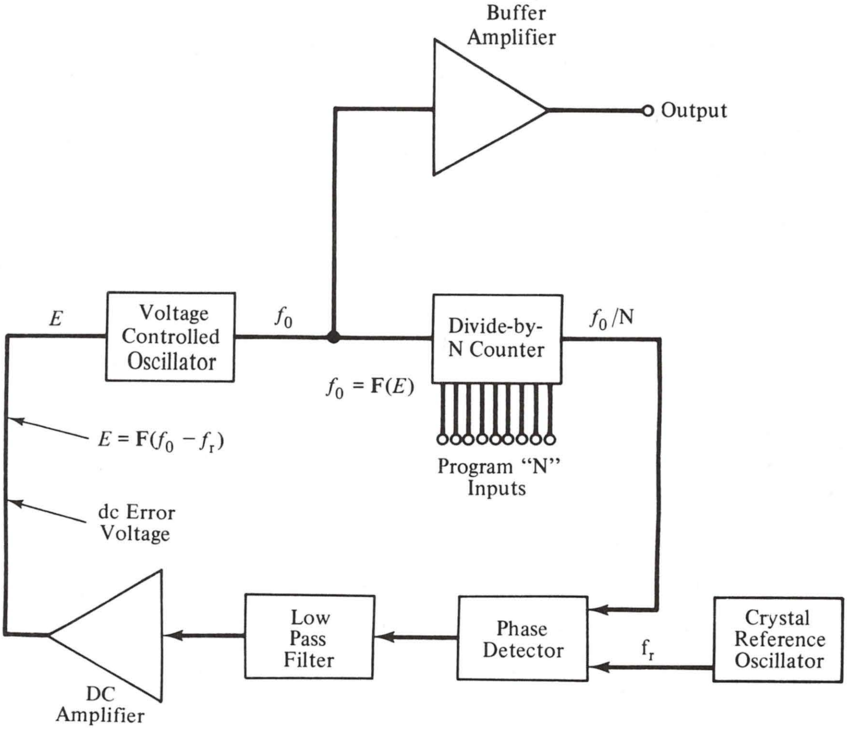


Figure 9-18. Block diagram of a PLL synthesizer.

problems that made them essentially either useless or too costly. In one technique, for example, a wide-band frequency spectrum is created. RF filters tuned to the various harmonics and frequency components of the spectrum pick off the components required to construct any given frequency. Most recently, the "crystalplexer" synthesizers have become used in some equipment. In these instruments a bank of crystal oscillators are heterodyned together to produce the desired frequency. In many 23 channel Citizen Band (CB) transceivers the crystalplexer synthesizer reduced the total number of crystals needed from 46 (i.e., two, one each in transmitter and receiver, for each channel of operation) to only 11 or 12. In that application, the crystalplexer was economic. But in applications using a wider frequency range the technique becomes prohibitively expensive. Modern wide-range frequency synthesizers use *phase locked loop* (PLL) circuits to generate the output frequencies.

A simplified block diagram of a PLL circuit is shown in Fig. 9-18. The principal components of a PLL circuit are a *voltage controlled oscillator* (VCO), *divide-by-N counter, phase detector, reference oscillator, low-pass filter,* and a *dc amplifier*.

The VCO oscillates at RF frequency F_0 that is a function of dc control voltage E. Voltage E is created by comparing the output frequency of the divide-by-N counter (a subharmonic of F_0) to a low-frequency reference oscillator signal. A phase detector circuit makes the comparison.

The divide-by-N counter is a digital counter circuit that divides the input frequency F_0 by an integer N. The integer division ratio is set by the binary word applied to the *program-N* inputs of the counter. The output of the counter will be a frequency that is equal to F_0/N.

The output of the N-counter and a low-frequency reference (i.e., 5 kHz to 10 kHz) are applied to opposite inputs of the phase detector circuit. As long as F_r is equal to (F_0/N), then the VCO will oscillate at the frequency determined by the dc amplifier offset control and the N-code applied to the counter. But if the VCO should drift, then the two signals applied to the phase detector are no longer equal, so the dc error voltage at the output of the phase detector changes in a direction and amount needed to pull the VCO back on frequency. The PLL, then, constantly corrects changes in the VCO frequency. The stability and precision of this type of signal generator, then, is essentially that of the crystal reference oscillator (which can be quite good).

Deliberate frequency changes can be made by reprogramming the divide-by-N counter to a different division ratio. The signal generator in Fig. 9-19 uses a numeric keyboard to enter the desired frequency. The operator enters the frequency, and the keyboard produces digital words that are interpreted as N-codes by the counter.

In most wide-range PLL synthesizers, several PLL circuits are used in a heterodyne system that is similar to the crystalplexer (although considerably more sophisticated). An example of such a signal generator is the Marconi Instruments Model TF2020, shown in block diagram form in Fig. 9-20.

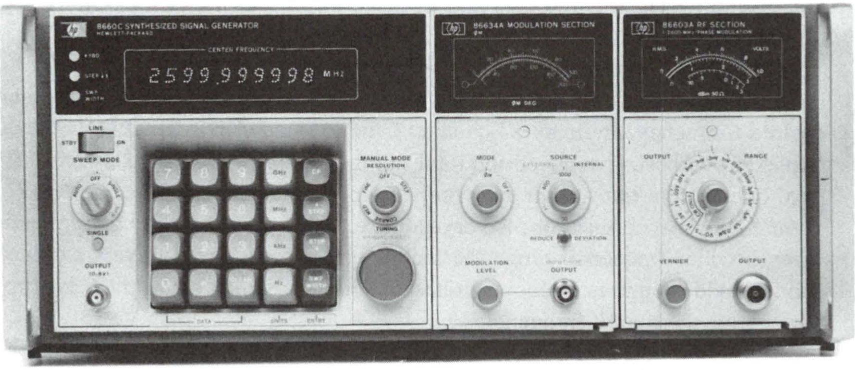


Figure 9-19. RF synthesizer-type signal generator. (Courtesy of Hewlett-Packard.)

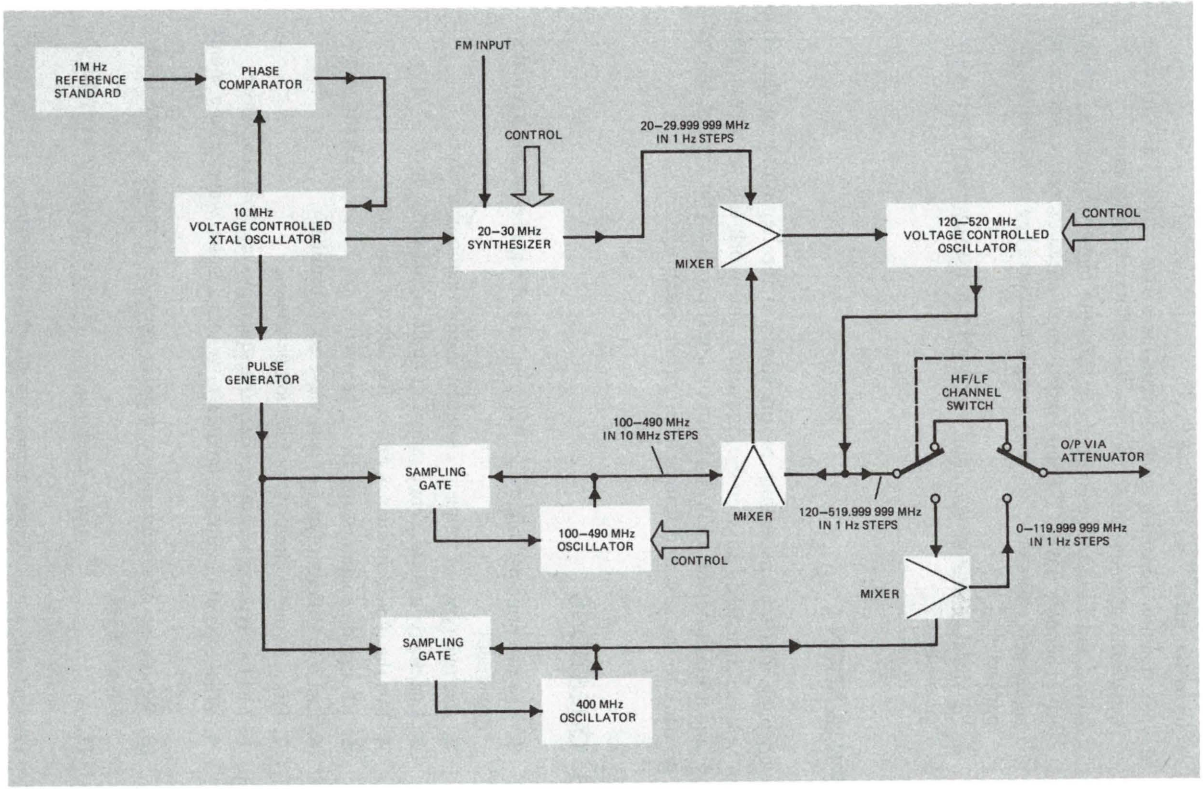


Figure 9-20. Block diagram of a wide range PLL-synthesized RF signal generator.

9-13 SUMMARY

1. Audio signal generators usually produce sine and square waves over the range of at least 20 Hz and 20 kHz. Some will operate down to 1 Hz and up to 100 kHz.
2. The function generator generally produces sine, square, and triangle waveforms at frequencies from fractional Hz to the mHz range.
3. Audio and function generators use 600-ohm output impedance, while RF generators have a 50-ohm output impedance. Proper operation of signal generators depends upon using a load matched to the output impedance.
4. RF signal generators operating over 1 gHz are usually called *microwave generators*.

9-14 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

9-15 QUESTIONS

1. List two principal differences between an audio generator and a function generator.
2. List two types of frequency selector used in audio generators.
3. List the typical frequency ranges of audio signal generators.
4. List the typical frequency ranges for function generators.
5. RF signal generators operate in the over- ______________-kHz region.
6. List four different types of sine wave audio oscillator.
7. Draw a simple circuit for a function generator.
8. List two types of circuit for creating square waves from sine waves.
9. What type of circuit can be used to generate a sine wave from a square wave?
10. What type of circuit can be used to generate a triangle wave from a square wave?
11. A ______________ audio generator uses two RF oscillators to create the audio signal.
12. List two ways to create a *linear* sawtooth waveform.
13. Some RF signal generators use a(n) ______________ to keep the output amplitude constant.
14. List the principal stages of a PLL synthesizer.

9-16 PROBLEMS

1. A standard RF signal generator is matched with its load. Calculate the voltage produced across the load if the generator attenuator is set to -10 dBm.
2. A standard audio generator is connected to a 600-ohm load. Find the rms voltage across the load if the signal generator is producing $+3$ VU. (*Hint:* 0 VU is 1 mV across 600 ohms, and VU is a power dB scale.)
3. A 10–15-mHz PLL operates from a 5-kHz reference. Find the N-code required to produce a VCO output frequency of 12.51 mHz when a divide-by-100 predivider is used ahead of the divide-by-N counter.

chapter 10

GRAPHICS RECORDING SYSTEMS

10-1 OBJECTIVES

1. To learn the principles of operation behind the PMMC galvanometer chart recorder.
2. To learn the principles behind the operation of servomechanism recording potentiometers.
3. To learn the causes of, and solutions for, certain types of recording problem.
4. To learn the different types of writing system used in chart recorders.

10-2 SELF-EVALUATION

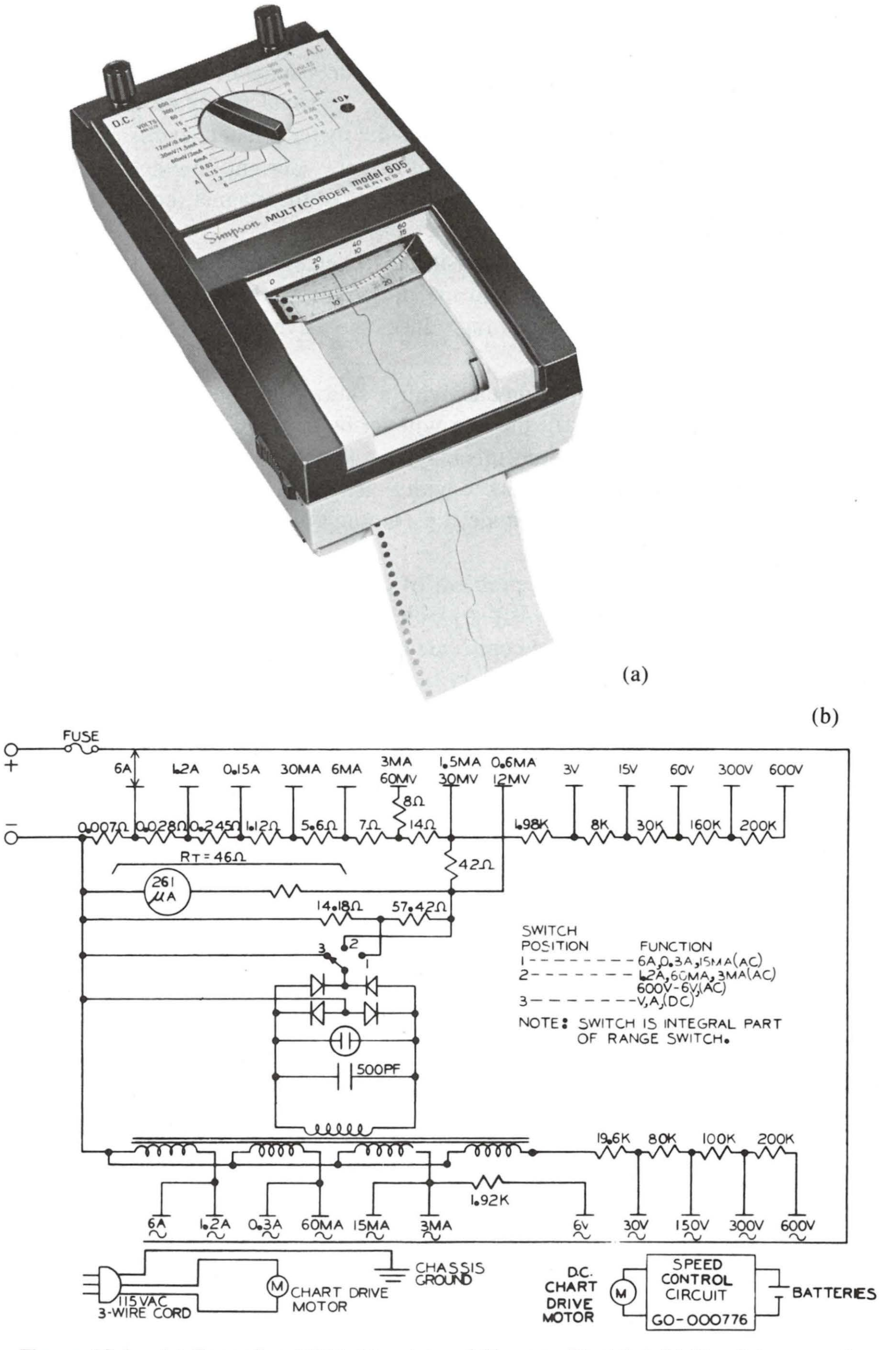
Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. Describe in your own words a PMMC galvanometer.
2. Describe the operation of a recording potentiometer.
3. What type of paper is used in thermal recorders?
4. What is a deadband signal?

10-3 THE RECORDING VOM

Figure 10-1 shows a recording volt-ohm-milliammeter (VOM); the actual instrument is shown in Fig. 10-1(a) and its circuit is shown in Fig. 10-1(b). The instrument is essentially a VOM configured so that its pointer rests over the chart paper. A cam on the paper drive motor lifts the pen up and down so that it taps the paper, leaving a trail of dots that follow the pointer position.

This type of recorder uses a very slow paper drive speed, up to about 12 inches per hour and as slow as 1 inch per hour. This type of instrument is especially useful for recording changes and trends occurring over relatively long periods of time.



(a)

(b)

Figure 10-1. (a) Recording VOM. (Courtesy of Simpson Electric.) (b) Circuit for recording VOM. (Courtesy of Simpson Electric.)

10-4 PMMC INSTRUMENTS

Many strip chart recorders use a *permanent magnet moving coil* (PMMC) galvanometer that is similar to the dc D'Arsonval meter movement discussed in Chapter 1. The writing pen replaces the meter pointer, and is attached to a moving coil in the field of a permanent magnet. Current flowing in the coil creates a magnetic field that interacts with the field of the magnet, and causes deflection of the coil-pen assembly.

The tip of the pen is positioned over a strip of chart paper that is pulled under the pen tip at a constant speed, thereby establishing a time base. Deflection of the pen across the paper reproduces the waveshape of the signal applied to the coil.

A PMMC galvanometer with a short pen arc will write in a *curvilinear* manner [Fig. 10-2(a)]; the pen will scribe an *arc* at its tip instead of a straight line. In some applications this may be tolerable, and the user merely records on a strip chart paper that is marked with a curved grid. This type of application would be found where the amplitude, i.e., amount of deflection, is more important than the waveshape.

A solution to the problem of curvilinear recording that is used in many higher-priced machines is the pivoted pen motor assembly as shown in Fig. 10-2(b). The PMMC is not connected directly to the pen, but through a mechanical linkage that translates the curvilinear motion of the moving coil to *rectilinear* motion at the pen's tip.

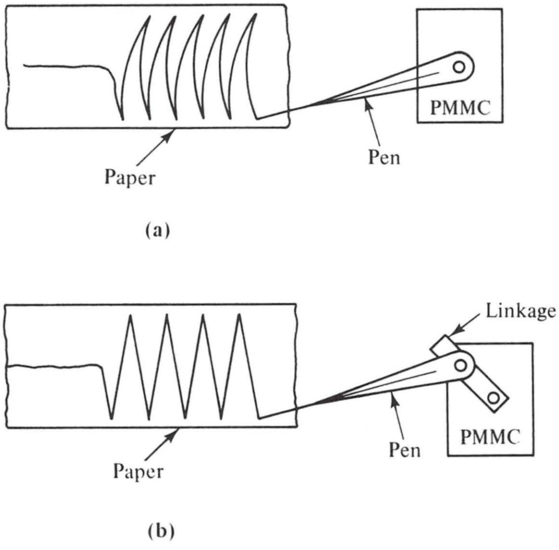
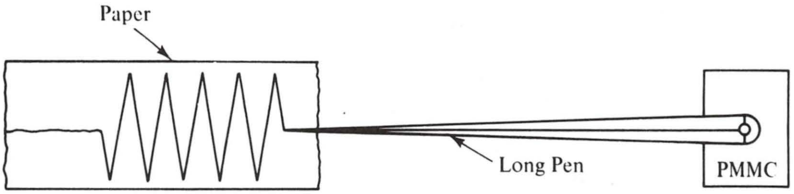
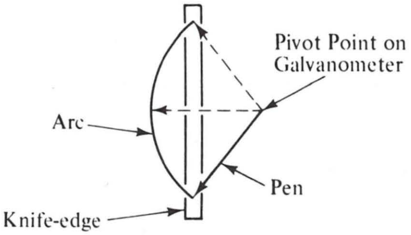


Figure 10-2. (a) Curvilinear recording. (b) Rectilinear recording.



(c) Pseudo-rectilinear Recording



(d)

Figure 10-2. *(cont.)* (c) Pseudorectilinear recording. (d) Knife edge creates rectilinear trace.

A pseudorectilinear technique is shown in Fig. 10-2(c). In this type of assembly the pen is very long compared with the width of the chart paper. The pen tip, therefore, travels in an arc whose length is very small relative to the radius, i.e., pen length. The trace, then, is nearly linear. Some curvature will exist, but for many applications it is tolerable.

There are several writing methods used in strip chart recorders, but two are amenable to a special type of pseudorectilinear recording: *thermal* and *direct contact*. Both of these writing systems can use a *knife edge,* also called a *writing edge,* to effectively linearize the tracing [see Fig. 10-2(d)]. The pen tip still travels in a curvilinear manner, but the trace is very nearly rectilinear because the knife edge is straight. This technique works in thermal and direct pressure systems because the "pen," which is actually a *stylus,* can write anywhere along its length. By keeping a straight knife edge in a fixed position beneath the stylus, and allowing the stylus's point of contact with the paper to vary, we obtain a rectilinear recording. It is the straightness of the knife edge that creates the rectilinear operation.

10-5 PMMC WRITING SYSTEMS

There are several different writing systems used on PMMC recorders: *direct pressure, thermal ink pen, ink jet,* and *optical.*

Those mechanisms that use pens or styli, i.e., direct pressure, thermal or ink pen, tend to have a relatively low-frequency response due to the inertia of the massive pen or stylus. Most such recorders have an upper −3 dB point in the response of 100 to 200 Hz. Ink jet and optical types, on the other hand, use lighter weight fixtures, so have a higher frequency response. An upper −3 dB point of 1000 Hz is usual, and up to about 2000 Hz is quite possible.

The direct pressure writer uses a special paper treated so that it has a carbonized back. When pressure is applied to the front surface of the paper a black mark appears at the pressure point. The Simpson 605 shown in Fig. 10-1 uses a pressure writing system. In that instrument the meter pointer taps the paper at the point where a mark is required.

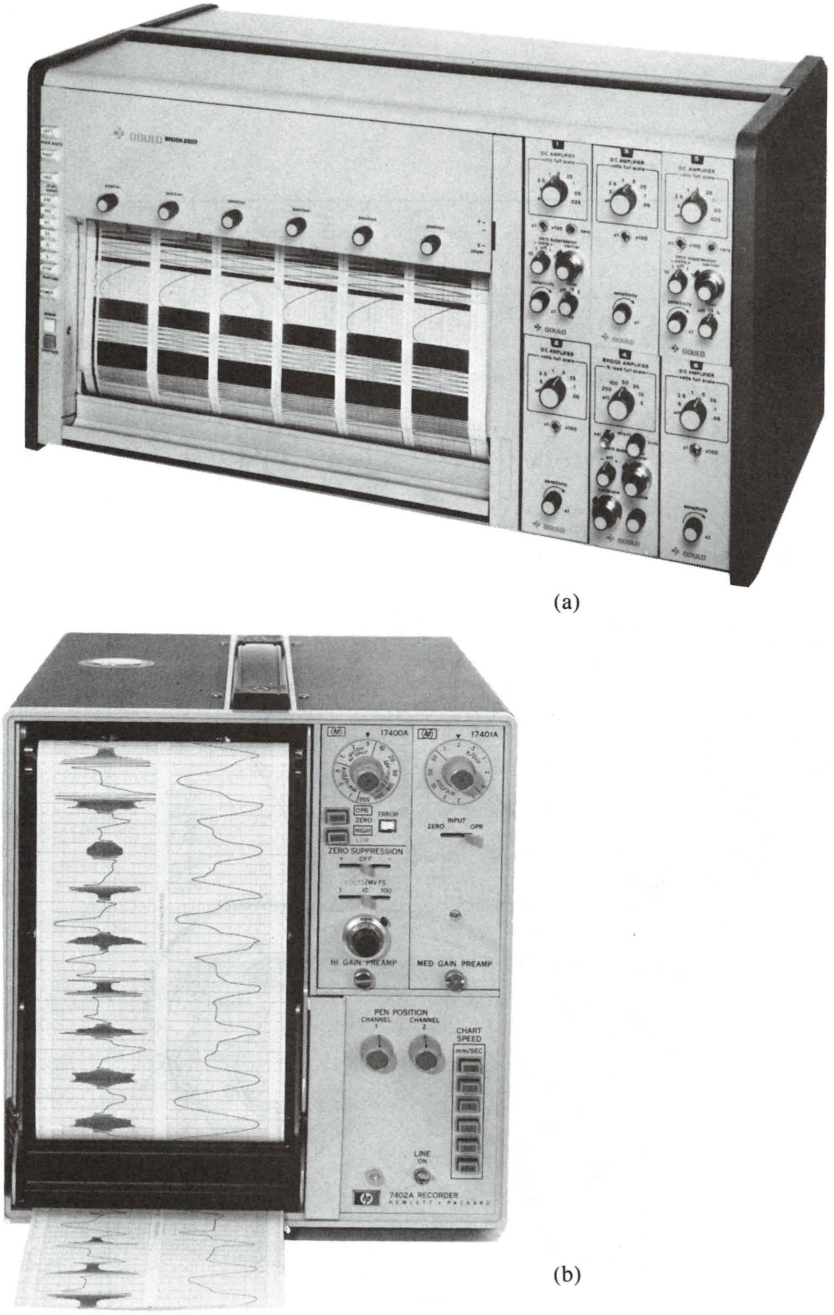
The thermal recording system also uses a special paper, but in that type of recorder the paper is *paraffin* treated so that it turns black when it is heated.

The writing instrument in a thermal recorder is a *heated stylus.* Early models formed the stylus tip from a U-shaped electrical resistance element, while modern models use a resistance element inside a cylindrical metal stylus. In both cases a low-voltage electrical current is passed through the element, heating almost to incandescence. The heated stylus leaves a black mark on the treated paper wherever it touches.

Ink pen systems write using a hollow pen and ink. In some machines the ink is relatively light, and pressure is applied manually through an atomizer type of pump (i.e., squeeze ball). Other machines, such as the Gould-Brush instruments [see Fig. 10-3(a)] and Hewlett-Packard Model 7402 [Fig. 10-3(b)], use a thick, viscous ink. The pressure is supplied (see Fig. 10-4) by a spring-driven pistol inside the ink cartridge and an ink manifold that distributes ink to the several pens. Pressure is applied to the manifold by a solenoid that is energized when the machine is turned on.

A related type of writing system, used in some high-frequency response European models, is the high-velocity ink jet. In that type of recorder ink under high pressure is fed to a nozzle mounted on the PMMC galvanometer in place of the pen. The ink jet is directed at the paper, and, when the system is properly adjusted, produces a line that is almost as fine as that produced by thermal and pen recorders. Only a small amount of trace fuzziness is apparent, and this is due to ink splattering.

The high-velocity ink jet recorder finds use where a high-frequency response is needed. The low mass of the ink jet nozzle, compared with the mass of ink pens and thermal styli, gives the recorder its characteristic 1- to 2-kHz frequency response.



(a)

(b)

Figure 10-3. (a) Multichannel direct writing recorder. (Courtesy of Gould/Brush.) (b) Two-channel recorder. (Courtesy of Hewlett-Packard.)

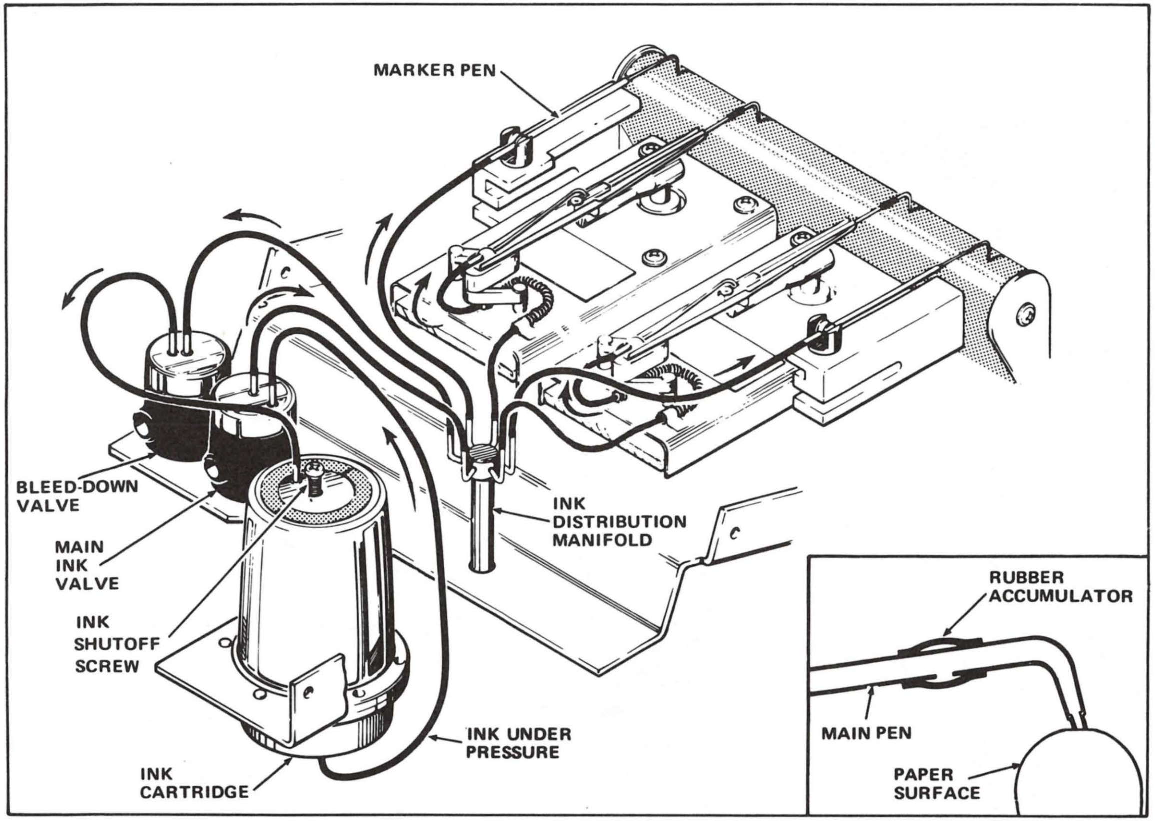


Figure 10-4. Pressurized ink system. (Courtesy of Hewlett-Packard.)

An example of an optical PMMC recorder is shown in Fig. 10-5(a), while a cathode ray tube (CRT) optical recorder is shown in Fig. 10-5(b).

The PMMC optical writer uses a *mirror* mounted on the galvanometer to reflect a light beam from a collimated light source onto the photosensitive chart paper. In most such machines the paper is at least six inches wide, so a greater span, or resolution, is possible. Additionally, on a multichannel recorder, the time relationships between two different traces can be more easily recognized be-

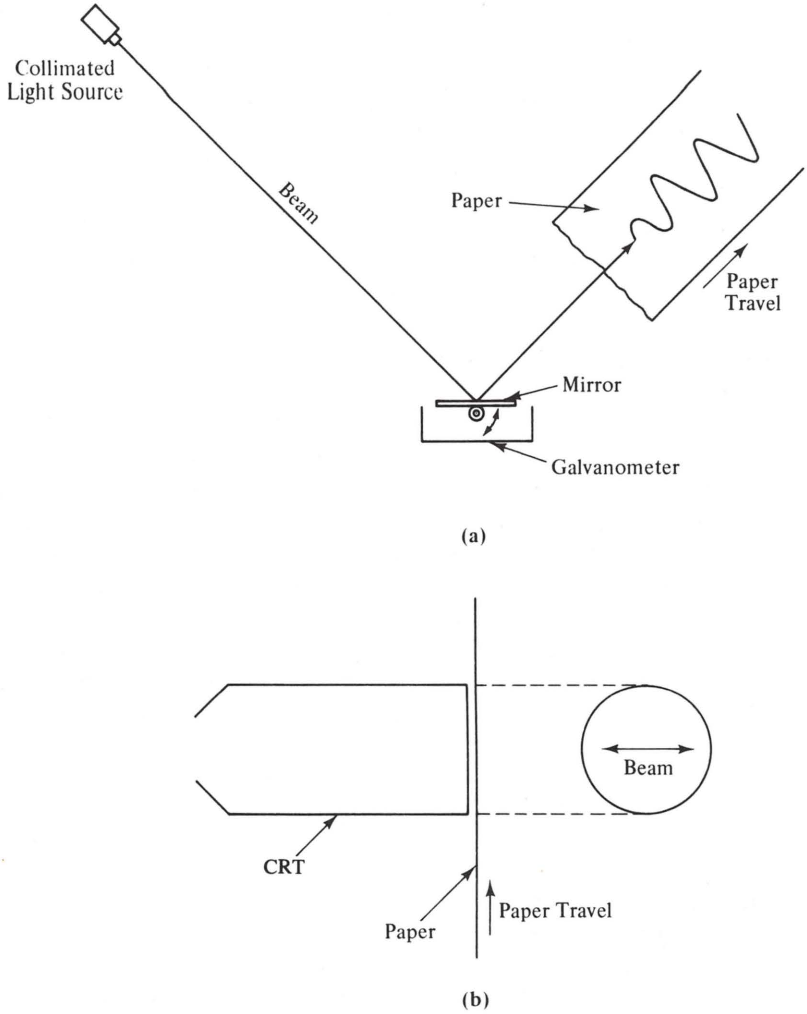


Figure 10-5. (a) Galvanometer/mirror-type optical recorder. (b) CRT camera optical recorder.

cause the beams can be made to *cross* each other—a difficult design trick with mechanical pens and styli.

The paper is exposed to an ultraviolet light that develops the trace as the paper comes out of the recorder. The trace will fade over a long period of time, however, unless the paper is stored in a light-tight box, or unless it is wet-developed in the manner of any photographic printing paper.

A few models use a cathode ray tube, or CRT, as in Fig. 10-5(b). A leading example is the VR-6 and VR-12 recorders by *Electronics-for-Medicine*. This type of strip chart recorder is called a *camera recorder*.

There is no time-base sweep on the CRT screen; the beam sweeps back and forth in the Y-direction only in response to amplitude variations of the input signal [examine Fig. 10-5(b)]. A time base is provided by the speed at which the paper is passed across the CRT screen.

The frequency response of the PMMC optical recorder is better than that of any other PMMC system because of the low mass of the reflecting mirror. The CRT camera, however, has an even higher frequency response, because it is limited only by the photographic writing speed of the recording paper.

10-6 RECORDING POTENTIOMETERS AND SERVORECORDERS

In Chapter 4 we discussed the potentiometer method for making comparison measurements. A three-terminal variable resistor is connected to produce an output voltage that is a function of a reference voltage and the *position* of the variable resistor's wiper arm. A galvanometer meter connected between the wiper and an unknown voltage will read *zero* when the known voltage from the potentiometer is equal to the unknown voltage.

In our Chapter 4 example the potentiometer had to be manually nulled, but it is possible to a build a *self-nulling* potentiometer. If a pen is connected to the potentiometer's wiper drive mechanism, then the voltage can be recorded on paper.

Figure 10-6 shows the basic dc potentiometer servorecorder mechanism. The pen is attached to a string that is wound around two idler pulleys and a drive pulley on the shaft of a dc servomotor. The pen assembly is also linked with a potentiometer ($R1$) in such a way that the position of the wiper arm on the resistance element is proportional to the pen position.

The potentiometer element is connected across a precision, stable reference potential E_{ref}, so potential E will represent the position of the pen. That is to say, E is the electrical *analog* of pen position. When the pen is at the left-hand side of the paper (in Fig. 10-6), E is zero, and when the pen is full scale, i.e., at the right-hand side of the paper, E is equal to E_{ref}.

The pen position is controlled by the dc servomotor, which is in turn driven

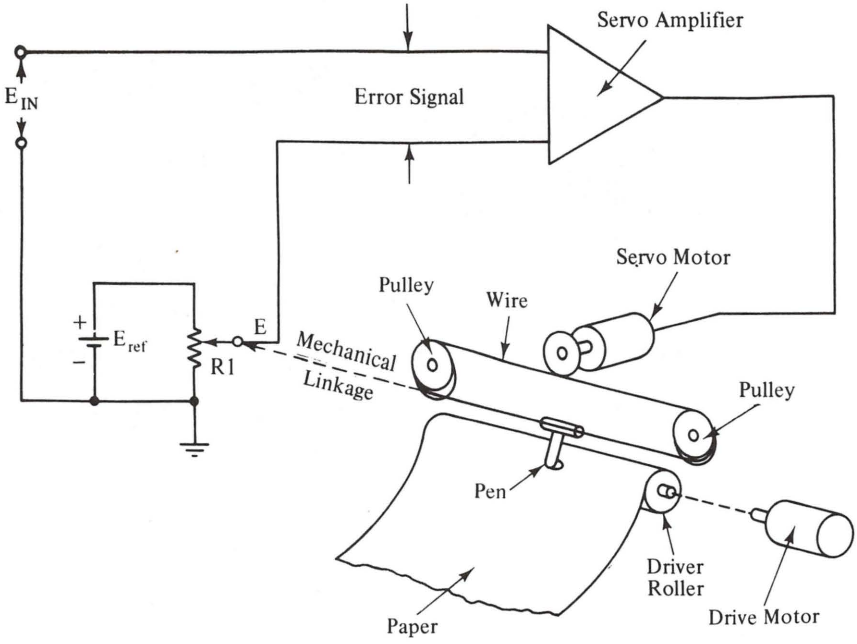


Figure 10-6. Servorecorder mechanism.

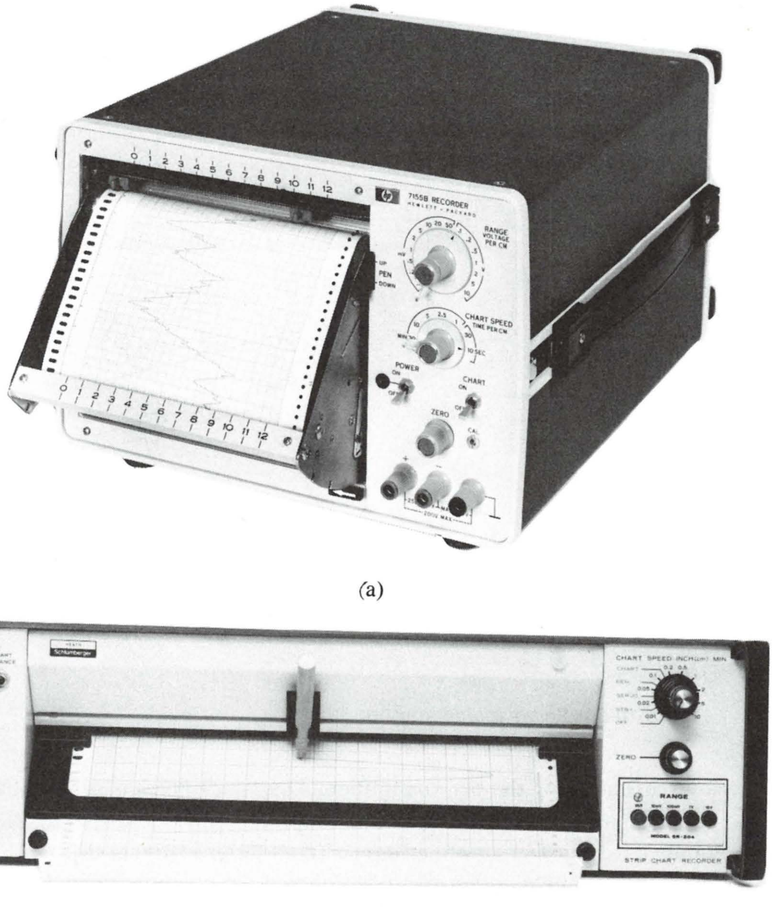
by the output of the servoamplifier. This amplifier has differential inputs. The input signal E_{in} is applied to one amplifier input, and the position signal E is applied to the other amplifier input. The difference signal ($E_{\text{in}} - E$) represents the *error* between the *actual* pen position and the position it *should* take in response to the command from E_{in}.

If the error signal is zero, then the amplifier output is also zero, so the motor remains turned off. But if $E \neq E_{\text{in}}$, then the amplifier sees an input signal and creates an output signal that turns on the motor.

The motor drives the pen-potentiometer assembly in such a direction as to *cancel* the error signal. When the input signal and position signal are equal, then the motor shuts off and the pen remains at rest.

A paper drive motor forms a time base because it pulls the paper underneath the pen at a fixed, constant rate. In most servorecorders a sprocket drive is used instead of the roller drive system popular in PMMC machines. The paper used in these machines (see Fig. 10-7) has holes along each margin to accept the sprocket teeth.

Most high-grade servorecorders use either a *stepper motor* that rotates only a few degrees for each pulse applied to its winding, or a continuously running motor that drives the paper sprocket through a gearbox or transmission.



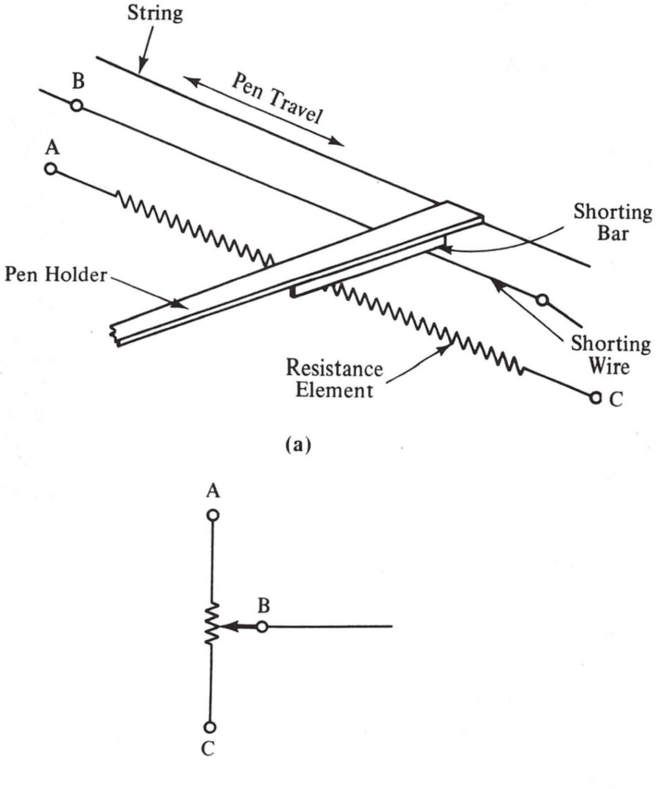
(a)

(b)

Figure 10-7. (a) Typical commercial servorecorder. (Courtesy of Hewlett-Packard.) (b) Another type of servorecorder. (Courtesy of The Heath Co.)

The stepper motor technique is low in cost, and can be very precise if a stable reference oscillator or the ac power mains are used as an electronic clock to develop the motor drive pulses. Integrated circuit digital logic circuits are often used to reduce the clock frequency to that frequency required to drive the motor at the desired speed.

The actual resistance element—or potentiometer if you prefer—used may be any of the following: slide-wire, rectilinear, or rotary models. The slide-wire system is often used, because it can be built with less friction and no mechanical linkage. An example of a slide-wire is shown in Fig. 10-8(a). A *resistance wire* and a *shorting wire* are stretched taut parallel to the direction of pen travel. A



(a)

(b)

Figure 10-8. (a) Slide-wire potentiometer in servorecorder. (b) Potentiometer.

shorting bar on the pen assembly serves as a wiper on the resistance element, and also serves to connect the shorting wire. The letters A, B, and C in Fig. 10-8(a) refer to the potentiometer terminals shown in Fig. 10-8(b). The shorting wire "B" serves as a terminal for the wiper.

10-7 X-Y RECORDERS AND PLOTTERS

An X-Y recorder (Fig. 10-9) uses two servomechanism pen assemblies mounted at right angles to each other. The vertical, i.e., Y-plane, amplifier moves the pen *vertically* along the *bar,* while the X-plane amplifier moves the bar back and forth across the paper.

The paper itself does not move, and is held in place either by clamps, or, in high-grade machines, by a vacuum pump that is connected to a hollow chamber

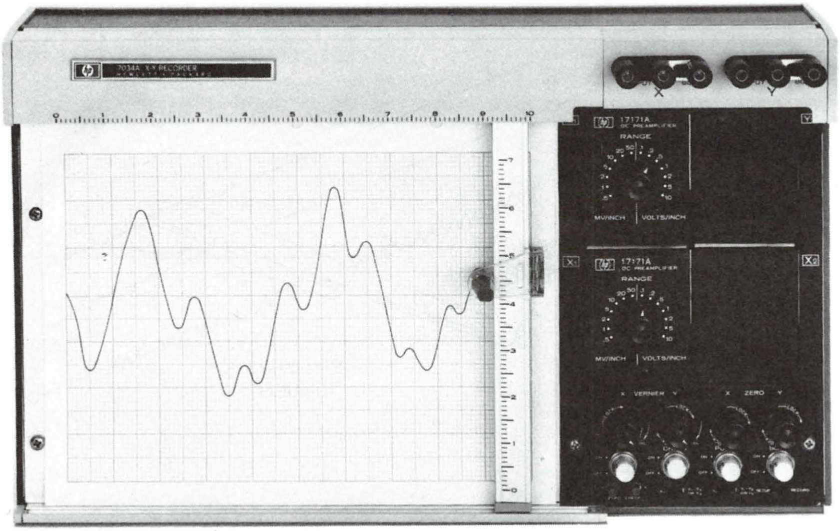


Figure 10-9. X-Y recorder.

just below the paper platform. Holes in the platform create the negative pressure needed to keep the paper in place.

One advantage to the X-Y recorder over either PMMC or servo strip chart recorders is that almost any type of paper may be used. Most X-Y recorders are designed to accept the standard sizes of graph paper preferred by scientists and engineers.

The X-Y recorder can be made to record time-varying signals by applying a linear *ramp* function to the X-channel input. The ramp is adjusted so that the bar traverses the horizontal width of the paper in the desired length of time.

An X-Y *plotter* (Fig. 10-10) is an X-Y recorder that has an electrically operated pen-lift mechanism. A plotter can make patterns of complex shape, including alphanumerics.

The X-Y plotter is often used as a computer peripheral device, and plots patterns under program control. The pen is lifted from the paper and repositioned in the correct spot for the line segments of the pattern to be created. The pen is then returned to the paper, and draws the desired segment. Each segment can be so drawn in its turn until the complete pattern is finished.

10-8 RECORDER PROBLEMS

The pen assembly always has a certain amount of inertia because of its mass, regardless of whether PMMC or servo systems are used. Because of its inertia, the pen assembly will not begin moving until a certain minimum signal is

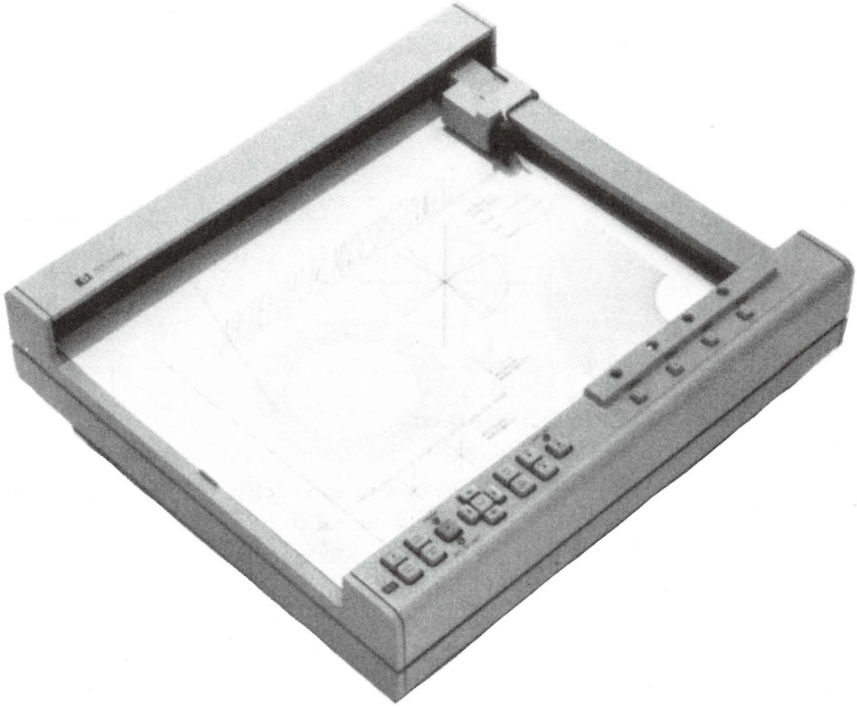


Figure 10-10. Graphics plotter.

applied. Figure 10-11 shows this *deadband* phenomenon. The definition of the deadband is "the largest amplitude signal to which the instrument will not respond." In most quality instrumentation recorders the deadband is 0.05 to 0.1 percent of full scale.

The deadband can cause distortion of the recorded trace, especially on low-amplitude signals whose level approximates the deadband itself. The solution to the problem of deadband is to slew through the deadband as rapidly as possible. That is to say, add sufficient preamplification to use as much of the recorder's span as possible.

Another problem is *overshoot* and *undershoot* in response to a step function. This problem also affects both PMMC and servorecorders. Figure 10-12 shows three types of response. In the properly damped recorder the pen will rise to the correct position with very little rounding of the trace. This situation is called *critical damping*.

A *subcritically damped* system will overshoot the correct position, and then hunt back and forth across the position for a few cycles before it settles on the correct point. If a square wave is applied to such a system, then the recording will appear to be *ringing*.

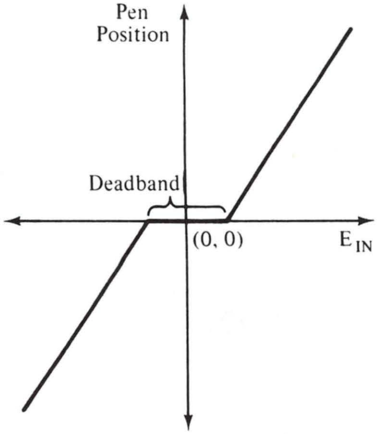


Figure 10-11. Principle of "deadband."

An *overcritically damped* system approaches the final value very slowly. The pen in that type of system will appear to be very sluggish. A square wave applied to such a system will have the rounded edges characteristic of passing the square wave through a low-pass filter.

The three damping factors are a mechanical analog of frequency filtering, and filters can be used to *correct* the damping problems. Some PMMC recorders have a *pen position transducer* inside the galvanometer housing, and servorecorders already have a position signal. The *position* signal can be integrated, i.e., low-pass filtered, to vary the damping of the pen assembly.

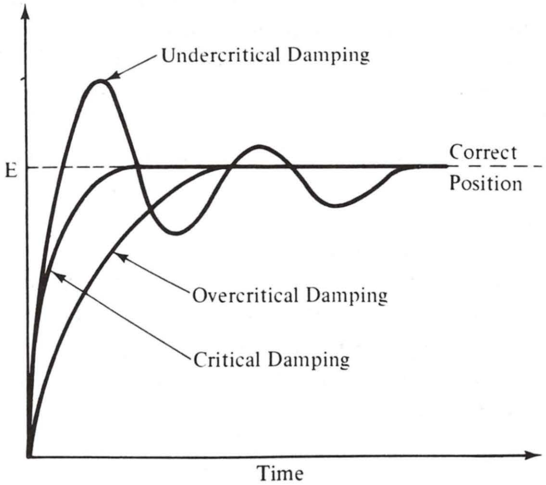


Figure 10-12. Effects of undercritical, overcritical, and critical damping.

In the PMMC recorder the integrated position signal is fed back to the servoamplifier to be summed with the error signal.

Pen assemblies can be damaged if they are allowed to strike the mechanical limits-of-travel stops. If a signal that is very large relative to the full-scale signal is applied to the input, then the pen may hit the mechanical stops and break.

Some PMMC recorder amplifiers use output limiting to guard against such damage. Alternatively, a pair of zener voltage regulator diodes connected back-to-back across the PMMC coil are sometimes used to accomplish the same job. The zener potential of the diodes is selected so that the diodes break over and conduct current *only* when a voltage greater than the normal full-scale potential is applied to the amplifier input.

If a PMMC position signal is available, then it can be *differentiated* to produce a *velocity* signal (i.e., $v = dx/dt$, from elementary physics), and the position signal may be differentiated twice to form an *acceleration* signal (i.e., $a = d^2x/dt^2 = dv/dt$). If appropriate circuitry is used, then these signals may be used to apply a hard braking signal to the PMMC coil if either pen velocity or acceleration exceeds certain limits. The pen will still strike the stops, but with considerably less force.

Most recorders have a *damping control* that is to be adjusted when a square wave, e.g., 1 Hz, is applied to the amplifier input. The damping control is adjusted to best "squareness" of the tracing.

10-9 PROBLEMS IN RECORDING

Recording strong signals, i.e., those in the over 100-mV range, is usually a straightforward job, and few problems exist. But as the full-scale value of signal amplitudes drops, then certain acute problems begin to develop. Signals in the millivolt range will exhibit some of these problems, while signals in the microvolt range invariably prove difficult to record.

Noise signals mixed with the signal will be recorded as a valid signal unless steps are taken to eliminate the noise, or at least reduce its value to a point of negligible effect. By *noise* we mean any electrical signal or tracing anomoly that is not part of the desired signal. Several different types of noise are recognized: *white noise, impulse noise,* and *interference noise.*

White noise supposedly contains all frequencies, so gets its name from the analogy to white light, which contains all colors. Such noise is also called *Gaussian* noise, although in reality it is neither "white" nor "Gaussian" unless there are no bandwidth limitations present. True white noise has a bandwidth of dc to daylight, and beyond. In most instrumentation systems, however, there are bandwidth limitations present, so the noise is actually *pseudogaussian.* True Gaussian noise can be eliminated by low-pass filtering, because it integrates to zero, given sufficient time. Bandwidth limited noise, however, does not integrate to zero, but to a very low value. The effect of low-pass filtering, then, is not total.

An analogy of pseudogaussian noise is the *hiss* heard between stations on an FM receiver. Most such noise in instrumentation systems is due to *thermal* sources, and has an rms value of

$$E_{\mathrm{n}} = \sqrt{4KTBR} \tag{10-1}$$

where E_{n} = the rms value of the noise potential in volts (V)
K = Boltzmann's constant 1.38×10^{-23} joules per degree Kelvin (1.38×10^{-23} J/°K)
T = the temperature in degrees Kelvin (°K)
B = the bandwidth in hertz (Hz)
R = the resistance in ohms (Ω)

Example 10-1

Find the amplitude of the noise signal in a circuit at room temperature (i.e., 300°K) if the bandwidth is 10 kHz and the resistance is 100 kΩ.

SOLUTION

$$E_{\mathrm{n}} = (4KTBR)^{1/2} \tag{10-1}$$
$$E_{\mathrm{n}} = [(4)(1.38 \times 10^{-23}\ \mathrm{J/^\circ K})(300^\circ\mathrm{K})(10^4\ \mathrm{Hz})(10\Omega)]^{1/2}$$
$$E_{\mathrm{n}} = [(4)(1.38 \times 10^{-14})(300)]^{1/2}$$
$$E_{\mathrm{n}} = (1.66 \times 10^{-11})^{1/2}\ \mathrm{V} = \mathbf{4.1 \times 10^{-6}\ V}$$

In Example 10-1 a signal of 4.1 microvolts is created by nothing more than thermal agitation of molecules in a circuit. Although this signal may appear to have a very low amplitude, keep in mind that many signals recorded on oscilloscopes and graphics recorder instruments are also in the microvolt range. The human electroencephalograph (i.e., brain-wave potentials) recorder, for example, must be able to deal with signals with peak amplitudes in the 10- to 100-μ V range. In that type of application the 4.1-μ V noise signal represents a sizable artifact.

The answer to this problem is to keep the circuit impedances in the early stages—i.e., those stages that most of the gain follows—very low so that the *R* term in Eq. (10-1) is low. Additionally, the bandwidth of the recording amplifiers should be limited to the limits required to faithfully reproduce the input signal waveform. This tactic will reduce the bandwidth term in Eq. (10-1).

There are also several other types of noise peculiar to solid-state amplifiers: *shot, Johnson,* and *flicker* noise. In low-cost amplifiers these noise sources can add up to a significant amplitude. Although low-pass filtering offers relief, it is better to specify a low noise preamplifier when one is ordering a recorder for use with low-level signals.

Impulse noise is due to local electrical disturbances such as arcs, lightning bolts, electric motors, and so forth. Shielding of signal lines will help somewhat, as will low-pass filtering, but the best solution is to eliminate the noise at its *source*. Filtering, incidentally, is a two-edged sword, and it must be done prudently. Filtering tends to *broaden* pulses, and that can create even more problems than it cures.

Other electrical devices nearby can induce signals into the instrumentation system, the chief among these sources being the 60-hertz power mains. It is wise to use only *differential* amplifier inputs, because of their high common mode rejection ratio (see Chapter 12). Signals from the desired source can be connected *across* the two inputs, and so become a differential signal, while the 60-Hz ac interference tends to affect both inputs equally (so is common mode).

It is sometimes possible, however, to manufacture a differential signal from a common mode signal. There are two methods by which this phenomenon occurs, and *both* involve the improper use of shields.

One source of this problem is called the *ground loop,* and is shown in Fig. 10-13(a). This problem arises from the use of too many grounds. In this example

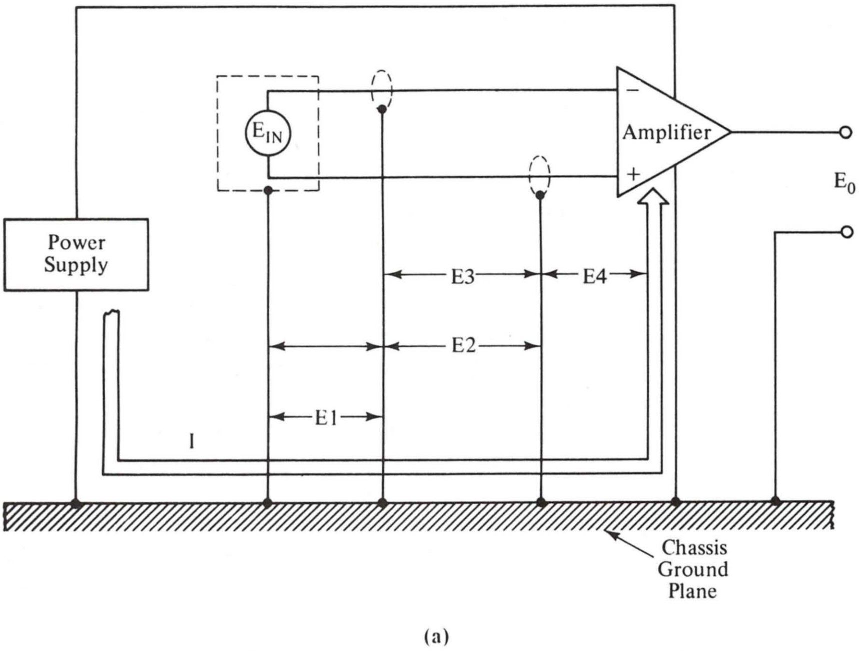
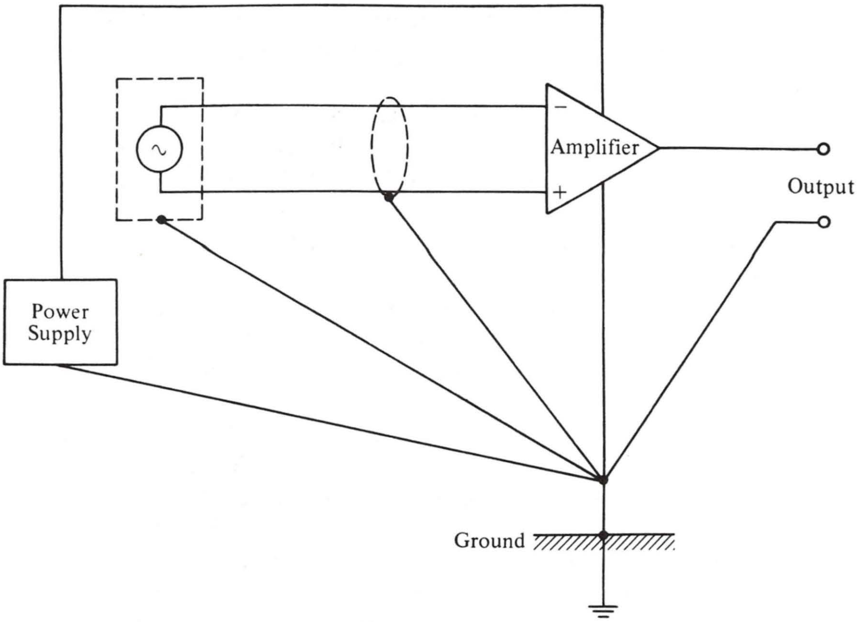


Figure 10-13. (a) Improper grounding.



(b)

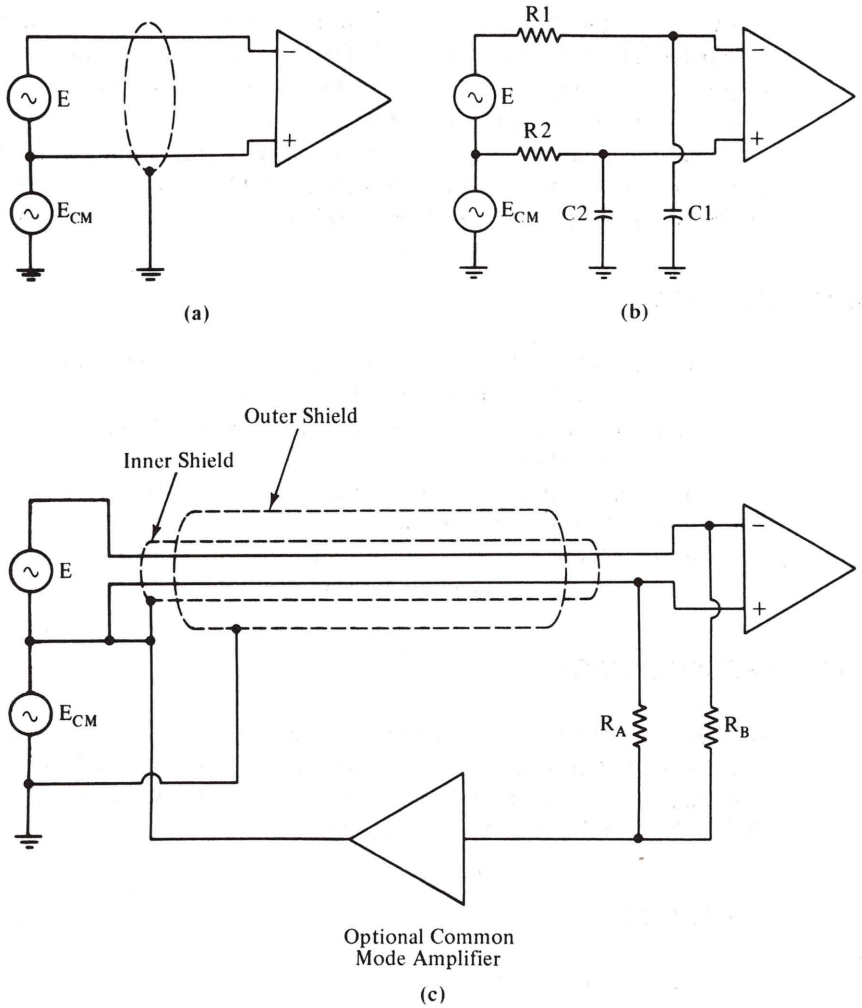
Figure 10-13. ***(cont.)*** (b) Proper, single-point grounding.

the shielded source, shielded input lines, the amplifier, and the power supply are all grounded to *different* points on the ground plane. Power supply dc currents *(I)* flow from the power supply at point A to the amplifier power common at point E. Since the ground has resistance, albeit very low resistance, the voltage drops $E1$ through $E4$ are formed. These are seen by the amplifier as valid signals, and can become especially troublesome if I is a varying current.

The solution to this problem is to use *single-point grounding,* as shown in Fig. 10-13(b). Some amplifiers used in sensitive graphic or CRT oscillographs keep the power and signal grounds separate, except at some single, specific common point. In fact, a few models go even further by creating several signal grounds, especially where both analog and digital signals might be encountered.

In some instances the shield on the input lines must *not* be grounded at *both* ends. In those cases it is usually better to ground the shield only at the amplifier end of the input cable.

Figure 10-14 shows some of the causes and cures for differential artifact signals manufactured from common mode sources. The circuit in Fig. 10-14(a) uses standard single shielding, but the equivalent circuit shown in Fig. 10-14(b) reveals a problem. The shield produces a capacitance to ground with the input wires ($C1$ and $C2$), and there are always cable ohmic and source resistances, represented here by $R1$ and $R2$. The system works well if $R1 = R2$ and $C1 = C2$, but



(a)

(b)

(c)

Figure 10-14. (a) Differential and common-mode signals. (b) Equivalent circuit. (c) Guard shield connections.

even small unbalances in the RC networks will allow E_{cm} to manufacture a differential signal. In that case, it is found that $E_{c1} \neq E_{c2}$, so the amplifier sees what it believes to be a valid input signal, which may be a dc offset or an ac signal such as the 60-Hz interference from the power mains.

A *guard shield* [Fig. 10-14(c)] circuit can be used to overcome this problem. The guard shield is driven by signals from the two input lines through high-value resistors R_a and R_b, and in many cases, a common mode amplifier.

This tactic has the effect of placing both sides of the cable capacitances at the same potential, so that $E_{c1} = E_{c2} = 0$. The outer shield is not strictly necessary, but it can serve to reduce coupling from the outside common mode source to the input lines, thereby reducing the value of E_{cm}.

It is sometimes difficult to predict all possible sources of input interference, so it often pays to try several different input shielding configurations to find the one that works best in the specific application.

10-10 SUMMARY

1. The PMMC galvanometer is analogous to a D'Arsonval meter movement in which the pointer is replaced by a pen.
2. The recording potentiometer uses a null-seeking servo system to drive the pen assembly.
3. Several writing systems are used: *direct pressure, thermal, pen and ink, ink jet, optical,* and *CRT camera.*
4. Most low-level signal recording problems can be solved by low-pass filtering and due regard for shielding.

10-11 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

10-12 QUESTIONS

1. Recording VOMs often use the ____________ writing system.
2. Describe the operation of a PMMC galvanometer. Draw a comparison between it and the D'Arsonval meter movement.
3. What is meant by *curvilinear* recording? Give any advantages and disadvantages to this system.
4. Describe a method by which a PMMC galvanometer can be constructed to render a *rectilinear* tracing.
5. Describe a pseudorectilinear PMMC recorder mechanism. Use a picture if you want it.
6. How does a *writing edge* (i.e., knife edge) render a rectilinear trace? What type of writing systems can use the writing edge?
7. Describe the following writing systems: *direct pressure* and *thermal.* Describe any special paper that is required in either of these systems.

8. Describe the following writing systems and compare their properties: *pen and ink* and *ink jet*.
9. Describe the *optical* PMMC galvanometer and *CRT camera* types of writing systems. Include in your discussion the dry and wet developing methods.
10. Use a block diagram to describe the self-balancing potentiometer recorder.
11. What is a *deadband?*
12. Describe *undershoot* and *overshoot*. How is it related to damping?
13. What is a *slide-wire potentiometer?* How is it constructed?
14. X-Y recorders use two ____________ recorder mechanisms positioned at right angles to each other.
15. How can X-Y recorders be made into Y-T (i.e., time) recorders?
16. The deadband on most quality recorders will be no more than _______ to _______ percent of full scale.
17. How can a *position* signal be used to adjust the damping of a PMMC circuit?
18. How may Gaussian noise be reduced?
19. Describe how a guard shield can reduce 60-Hz interference.
20. Describe how multiple-point grounding can create a differential component from a common mode signal.

10-13 PROBLEMS

1. Find the rms thermal noise potential generated by a 1-megohm resistor shunted by a 2-μF capacitor if the temperature is 25°C (i.e., room temperature).
2. A unity gan amplifier with a bandwidth of 200 kHz has a 10-megohm input resistance. If the amplifier is ideal, i.e., itself noiseless, and is operated at a temperature of 50°C, the thermal noise output potential will be ____________ V.

10-14 REFERENCES

1. Peter Strong, *Biophysical Measurements*. Tektronix, Inc. 1973.
2. Hewlett-Packard, *X-Y Recorder Applications,* Notes 1, 2, and 3.
3. James Diefenderfer, *Principles of Electronic Instrumentation*. Philadelphia, Pa.: W.B. Saunders, 1972.

chapter 11

LABORATORY AMPLIFIERS

11-1 OBJECTIVES

1. To learn the different forms of commercial amplifiers.
2. To learn how to specify and apply laboratory amplifiers.

11-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer a particular question, look for the answer as you read the text.

1. An isolation amplifier typically has an input impedance of_____ ohms.
2. Chopper amplifiers are used where _____ and noise are a problem because of high voltage gains.
3. An amplifier used for exciting a transducer with ac is called a _____ amplifier.
4. The lock-in amplifier is particularly useful when the signal is in the presence of a large amount of _____.
5. A differential voltage amplifier with a gain of 100 would be considered a _____ (low, medium, high) gain amplifier.

11-3 BASIC AMPLIFIERS

There are many applications for commercially built amplifiers. Such instruments are typically used in laboratories, industry, and field locations for data collection and distribution and other chores. Although it is quite possible for a skilled engineer to design amplifiers at substantially lower cost than the commercial equivalents, it is rarely cost-effective to do so given certain realities (time, compatibility, workload, etc.). In this chapter, we will examine some of the basic forms of amplifiers that can be purchased from instrument manufacturers.

Fig. 11-1 shows the three basic symbols used to represent amplifiers. These symbols are typically used for both commercial amplifiers that are whole and complete in themselves and also for integrated circuit operational amplifiers. The context in which the symbol is used will help you determine which is the case.

The standard single-ended amplifier symbol is shown in Fig. 11-1(a). This form of amplifier requires an input signal that is referenced to ground or a common bus; such an input is said to be *unbalanced*. Likewise, the output signal is also unbalanced. In some cases, the gain of the amplifier will be written inside of the symbol using a format such as "×100" (i.e., gain is times 100) or "$A_v = 100$" (which means voltage gain = 100). Generally, we cannot tell whether an amplifier such as that shown in Fig. 11-1(a) is inverting or noninverting unless additional information is provided. For example, an indication of "$-\times 100$" or "$-A_v = 100$" would denote an inverting amplifier. An inverting amplifier provides an output that is 180 degrees out of phase with the input signal, while a noninverting amplifier produces a signal that is in phase with its input signal.

The transfer function of an inverting amplifier is written:

$$V_o / V_{in} = -A_v \tag{11-1}$$

where V_o = the output signal voltage
V_{in} = the input signal voltage
A_v = the voltage gain factor of the amplifier

while the transfer function of a noninverting amplifier is the same, but without the minus sign in front of the gain symbol (A).

A differential amplifier symbol is shown in Fig. 11-1(b). This type of amplifier uses two inputs, inverting and noninverting. The inverting input produces an output that is 180 degrees out of phase with the input signal. The noninverting input, on the other hand, produces an output that is in phase with the

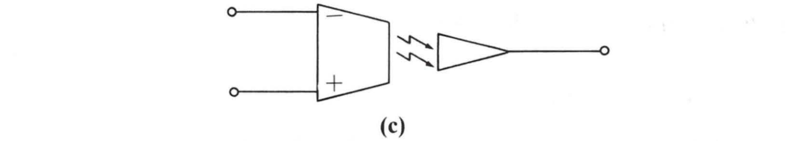


Figure 11-1. Amplifier symbols: (a) standard amplifier, (b) differential amplifier, (c) isolation amplifier.

input signal. Since both inputs see the same gain from the amplifier, an implication of applying the same signal to both inputs simultaneously is that the output is zero. This fact can be seen from the transfer equation for a differential amplifier:

$$V_o = (V1 - V2) \times A_{vd} \qquad (11\text{-}2)$$

where V_o = the output signal voltage
$V1$ = the signal voltage applied to the noninverting input
$V2$ = the signal voltage applied to the inverting input
A_{vd} = the differential voltage gain

The last generic type of amplifier symbol we will consider is the isolation amplifier, shown in Fig. 11-1(c). This type of amplifier isolates the input circuitry from the output circuitry. A typical hybrid isolation amplifier will have an input impedance (as measured to the ac power supply serving the nonisolated side) of more than 10^{12} ohms.

The principle applications for isolation involve safety of either people or instruments. For example, electrical shock considerations in biomedical amplifiers require the use of isolation amplifiers for signal acquisition. An electrocardiograph amplifier, for example, connects a patient to an ac-powered recording or display device. Since as little as 20 microamperes is considered dangerous to certain patients, leakage between the chassis and the ac line could place the patient at extreme risk. Thus, an isolation amplifier provides safety. We also see isolation amplifiers used when the environment, often a high voltage, would damage the recording circuitry. In biochemistry, for example, high voltages (1000 volts dc) are sometimes used for electrophoresis experiments. In this case, a constant current at high voltage will cause different proteins to migrate to different levels in a glass cylinder placed in the electrical field. If we needed to record temperature inside the electrophoretic solution, we would need to connect the thermocouple to the external recorders or data acquisition system via an isolation amplifier. Otherwise, the 1000 volt dc field would damage the electronic circuits.

11-4 AMPLIFIER CLASSIFICATIONS

Laboratory amplifiers are often classified according to gain. Typical categories are low gain, medium gain, and high gain. Although there is no official system of classification, and definitions tend to vary from one manufacturer to another, the following are generally followed:

Low gain. Voltage gain of 1 to 50.

Medium gain. Voltage gain of 50 to 500.

High gain. Voltage gain over 500.

We also categorize amplifiers as either ac or dc, or both. An ac amplifier will have a lower frequency response limit, and will not respond to dc signals. In the case of certain biophysical amplifiers, the lower −3dB point in the frequency response characteristic might be 0.05 Hz, but is not dc. The reason is that there are very low frequency components in these signals, but electrode drift problems argue against dc for the lower end frequency.

11-5 LABORATORY AMPLIFIER SYSTEMS

Laboratory amplifiers are often integrated into systems such as shown in Figs. 11-2 and 11-3. The Hewlett-Packard mainframe shown in Fig. 11-2 is one of several made by that company; it finds applications in a wide range of jobs from medicine to industrial signals acquisition. The amplifiers are in the form of plug-in modules. The mainframe accepts the plug-ins and provides them with mechanical support, dc power, ac carrier excitation (for transducer amplifiers), output signal routing,

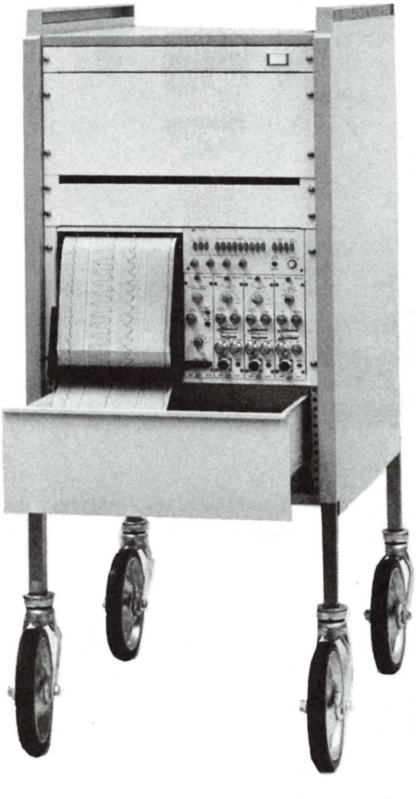


Figure 11-2. Laboratory instrumentation system by Hewlett-Packard.

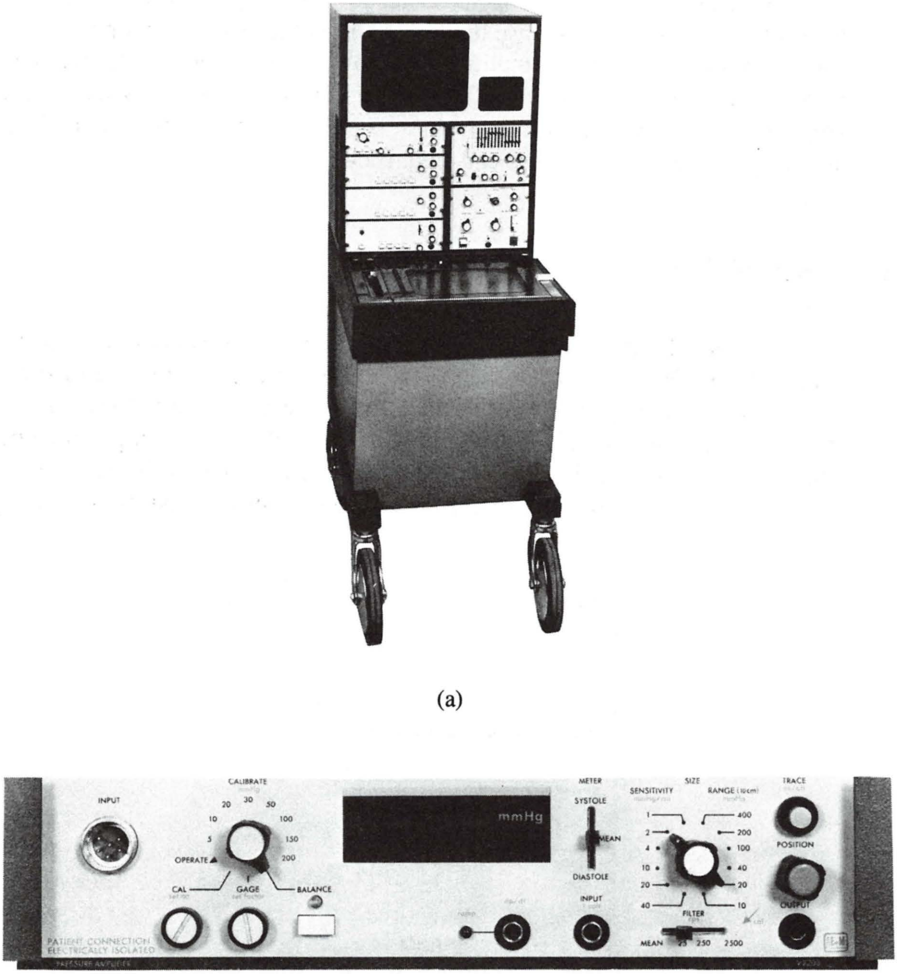


Figure 11-3. (a) Laboratory instrumentation system using a recording camera. (Courtesy of Electronics for Medicine, Inc.) (b) Plug-in pressure amplifier.

and (sometimes) input signal routing. In most cases, the input signals are applied through the front panel of each amplifier, although that is not universally true. This particular H-P configuration has the mainframe as part of a four-channel analog pen recorder that makes a permanent record of the waveforms and is mounted in a portable roll-around cart.

Fig. 11-3(a) shows another form of instrument package, this one by Electronics for Medicine. In this case, the plug-in amplifiers (example in Fig. 11-3(b)) are mounted in a housing with a long persistence analog CRT oscilloscope. The lower unit (horizontal) is a CRT camera recorder that makes copies of the trace appearing on the upper CRT. The plug-in shown in Fig. 11-3(b) is a blood pressure amplifier, so the output meter is calibrated in pressure units (millimeters of mercury, mmHg). Note that the gain control on the right is also calibrated in mmHg.

11-6 CHOPPER AMPLIFIERS

DC amplifiers have a certain inherent *drift* and tend to be *noisy*. These factors are not too important in low- and medium-gain applications (that is, gains less than 500), but loom very large indeed at high gain. For example, a 50-μ V/°C drift figure in an × 100 amplifier produces an output voltage of

$$(50\ \mu\text{V}/^\circ\text{C}) \times 100 = 5\ \text{mV}/^\circ\text{C} \tag{11-3}$$

which is tolerable in most cases. But in an ×100,000 amplifier, the output voltage would be

$$(50\ \mu\text{V}/^\circ\text{C}) \times 10^5 = 5\ \text{V}/^\circ\text{C}$$

and that amount of drift will probably obscure any real signals in a very short time.

Similarly, noise can be a problem in high-gain applications, where it had been negligible in most low- to medium-gain applications. Operational amplifier noise is usually specified in terms of *nanovolts per square root* hertz [i.e., $\text{noise}_{\text{rms}} = \text{nV(Hz)}^{1/2}$]. A typical low-cost operational amplifier has a noise specification of 100 $\text{nV(Hz)}^{1/2}$, so at a bandwidth of 10 kHz the noise amplitude will be

$$\text{Noise}_{\text{rms}} = 100\ \text{nV}\ (10^4\ \text{Hz})^{1/2}$$
$$\text{Noise}_{\text{rms}} = 100\ \text{nV}\ (10^2)$$
$$\text{Noise}_{\text{rms}} = 10^4\ \text{nV} = 10^{-5}\ \text{V}$$

In an ×100 amplifier, without low-pass filtering, the output amplitude will be only 1 mV, but in an ×100,000 amplifier it will be 1 volt.

A circuit called a *chopper amplifier* will solve both problems, because it makes use of an ac-coupled amplifier.

The drift problem is cured because of two properties of ac amplifiers; one is the inability to pass low-frequency (i.e., near-dc) changes such as those caused by drift, and the other is the ability to regulate the stage through the use of heavy doses of negative feedback.

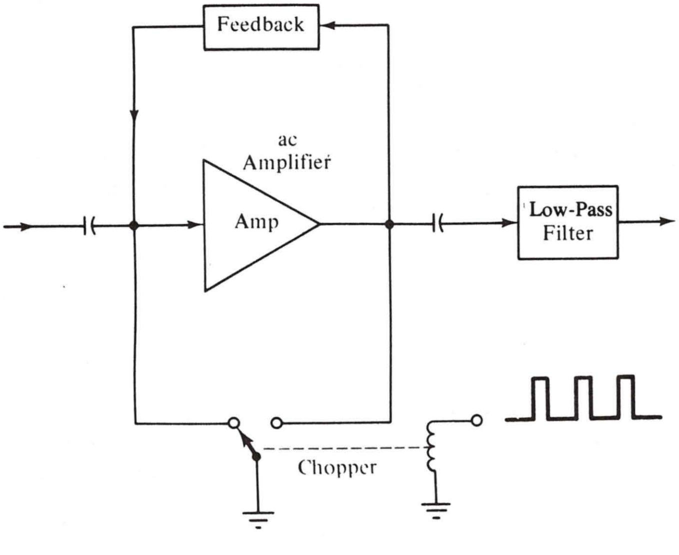
But many low analog signals are very low frequency, i.e., in the dc to 30 Hz range, so will not pass through such an amplifier. The answer to this problem is to *chop* the signal so that it passes through the ac amplifier, and then to demodulate the amplifier output signal to recover the original waveshape, but at a higher amplitude.

Figure 11-4(a) shows the basic chopper circuit. The traditional chopper is a vibrator-driven SPDT switch (*S*1) connected so that it alternately grounded first the input and then the output of the ac amplifier. An example of a chopped waveform is shown in Fig. 11-4(c). A low-pass filter following the amplifier filters out any residual chopper hash and any miscellaneous noise signals that may be present.

Most of these mechanical choppers use a chop rate of 400 Hz, although 60-, 100-, 200-, and 500-Hz choppers are also known. The main criterion for the chop rate is that it be twice the highest component frequency that is present in the input waveform. In other words, it must obey Nyquist's criterion.

A differential chopper amplifier is shown in Fig. 11-4(b). In this circuit an input transformer with a center-tapped primary is used. One input terminal is connected to the transformer center-tap, while the other input terminal is switched back and forth between the two ends of the primary winding.

A synchronous demodulator following the ac amplifier detects the signal, and restores the original, but now amplified, waveshape. Again, a low-pass filter smoothes out the signal.



(a)

Figure 11-4. (a) Chopper amplifier.

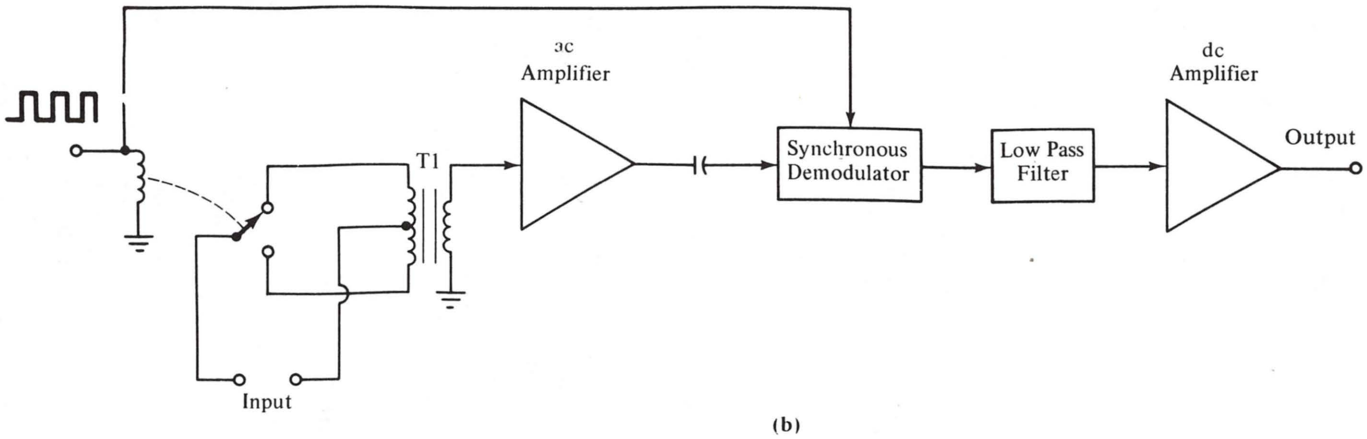
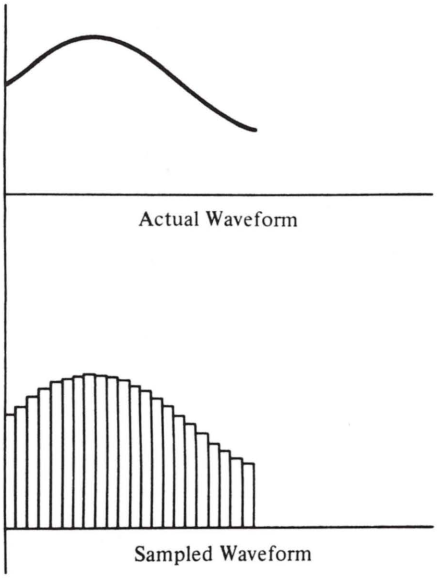


Figure 11-4. *(cont.)* (b) Differential chopper amplifier.



(c)

Figure 11-4. ***(cont.)*** (c) Continuous-vs-sampled analog waveforms.

The modern chopper amplifier may not use mechanical vibrator switches as the chopper. A pair of CMOS or JFET electronic switches driven out of phase with each other will perform the same job. Some monolithic or hybrid function module chopper amplifiers use a varactor switching bridge for the chopper.

The chopper amplifier limits noise because of the low-pass filter, and because the amplifier can have a narrow bandpass centered around the chopper frequency.

11-7 CARRIER AMPLIFIERS

A *carrier amplifier* is any type of signal-processing amplifier in which the signal carrying the desired information is modulated onto another signal, i.e., a "carrier." The chopper amplifier is considered by many to fit this definition, but is usually regarded as a type in its own right. The two principal carrier amplifiers are the dc-excited and ac-excited varieties.

Figure 11-5 shows a dc-excited carrier amplifier. The Wheatstone bridge transducer is excited by dc potential E. The output of the transducer, then, is a small dc voltage that varies with the value of the stimulating parameter. The transducer signal is usually of very low amplitude, and is noisy. An amplifier builds up the amplitude, and a low-pass filter removes much of the noise. In

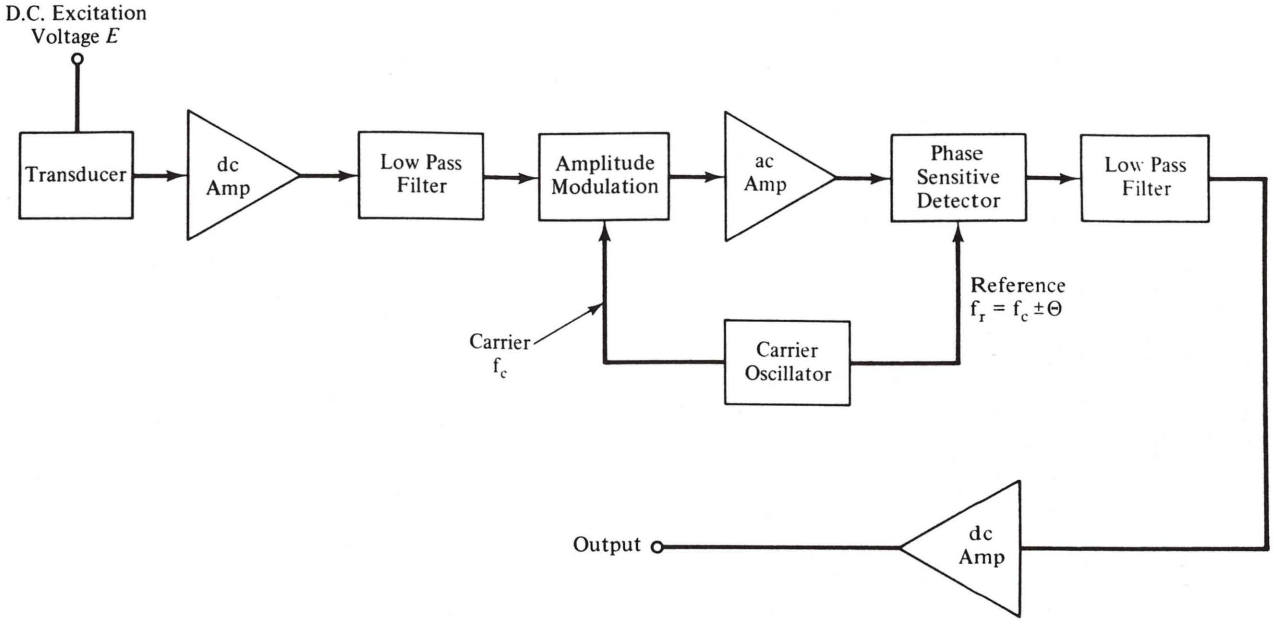


Figure 11-5. DC-excited carrier amplifier.

some models the first stage is actually a composite of these two functions, being a filter with gain.

The signal at the output of the amplifier-filter section is used to amplitude-modulate a carrier signal. Typical carrier frequencies range from 400 Hz to 25 kHz, with 1 kHz and 2.5 kHz being very common. The signal frequency response of a carrier amplifier is a function of the carrier frequency, and is usually considered to be one-fourth of the carrier frequency. A carrier or 400 Hz, then, is capable of signal frequency response to 100 Hz, while the 25-kHz carrier will support a frequency response of 6.25 kHz. Further amplification of the signal is provided by an ac amplifier.

The key to the performance of any carrier amplifier worthy of the name is the phase sensitive detector (PSD) that demodulates the amplified ac signal. Envelope detectors, while very simple and of low cost, suffer from an inability to discriminate between the real signal and spurious signals.

Figure 11-6 shows a simplified PSD circuit. Transistors *Q*1 and *Q*2 provide a return path to ground for the opposite ends of the secondary winding of input transformer *T*1. These transistors are alternately switched into an out of conduction by the reference signal in such a way that *Q*1 is off when *Q*2 is on, and vice versa. The output waveform of the PSD is a full-wave rectified version of the input signal.

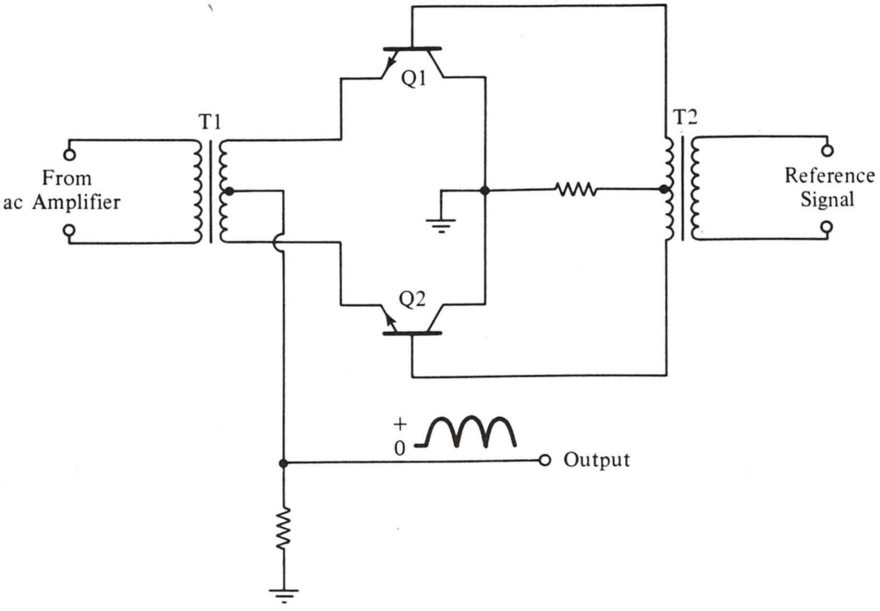


Figure 11-6. Phase sensitive detector.

Other electronic switching circuits are also used in PSD design. All systems are designed using the fact that a PSD is essentially an electronic DPDT switch. The digital PSD circuit most often seen uses a CMOS electronic IC switch such as the CD4016/CD4066. These switches are toggled by the reference frequency in such a way that the output is always positive-going, regardless of the phase of the input signal.

The advantages of the PSD include the fact that it rejects signals not of the carrier frequency, and certain signals that *are* of the carrier frequency. The PSD, for example, will reject even harmonics of the carrier frequency, and those components that are out of phase with the reference signal. The PSD will, however, respond to odd harmonics of the carrier frequency. Some carrier amplifiers seem to neglect this problem altogether. But in some cases, manufacturers will design the ac amplifier section to be a bandpass amplifier with a response limited to $F_c \pm (F_c/4)$. This response will eliminate any third, or higher-order, odd harmonics of the carrier frequency before they reach the PSD. It is then necessary only to assure the purity of the reference signal.

An alternate, but very common, form of carrier amplifier is the ac-excited circuit shown in Fig. 11-7. In this circuit the transducer is ac-excited by the carrier signal, eliminating the need for the amplitude modulator. The small ac signal from the transducer is amplified and filtered before being applied to the PSD circuit. Again, some designs use a bandpass ac amplifier to eliminate odd-harmonic response. This circuit allows adjustment of transducer offset errors in the PSD circuit instead of in the transducer, by varying the phase of the reference signal.

11-8 LOCK-IN AMPLIFIERS

The amplifiers discussed so far in this chapter produce relatively large amounts of noise, and will respond to noise present in the input signal. They suffer from shot noise, thermal noise, H-field noise, E-field noise, ground loop noise, and so forth. The noise voltage or power at the output is directly proportional to the square root of the circuit bandwidth. The *lock-in amplifier* is a special case of the carrier amplifier idea in which the bandwidth is very narrow. Some lock-in amplifiers use the carrier amplifier circuit of Fig. 11-7, but use an input amplifier with a very high Q bandpass. The carrier frequency may be anything between 1 Hz and 200 kHz. The lock-in principle works because the information signal is made to contain the carrier frequency in a way that is easy to demodulate and interpret. The ac amplifier accepts only a narrow band of frequencies centered about the carrier frequency. The narrowness of the bandwidth, which makes possible the improved signal-to-noise ratio, also limits the lock-in amplifier to very low-frequency input signals. Even then, it is sometimes necessary to time-average the signal for several seconds to obtain the needed data.

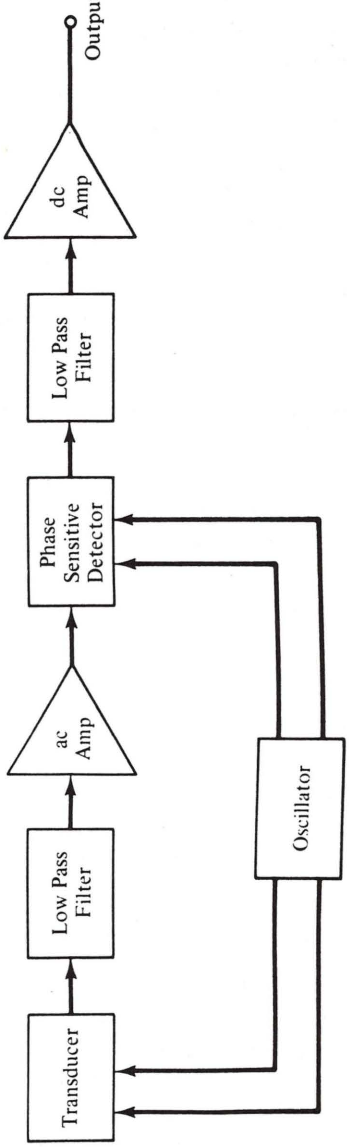


Figure 11-7. AC-excited carrier amplifier.

Lock-in amplifiers are capable of thinning out the noise, and retrieving signals that are otherwise "buried" in the noise level. Improvements of up to 85 dB are relatively easily obtained, and up to 100 dB is possible if the cost is no factor.

There are actually several different forms of lock-in amplifier. The type discussed above is perhaps the simplest type. It is merely a narrow band version of the ac-excited carrier amplifier. The lock-in amplifier of Fig. 11-8, however, uses a slightly different technique. It is called an autocorrelation amplifier. The carrier is modulated by the input signal, and then integrated (i.e., time-averaged). The output of the integrator is demodulated in a product detector circuit. The circuit in Fig. 11-8 produces very low output voltages for input signals that are not in phase with the reference signal, but produces a relatively high output at the proper frequency.

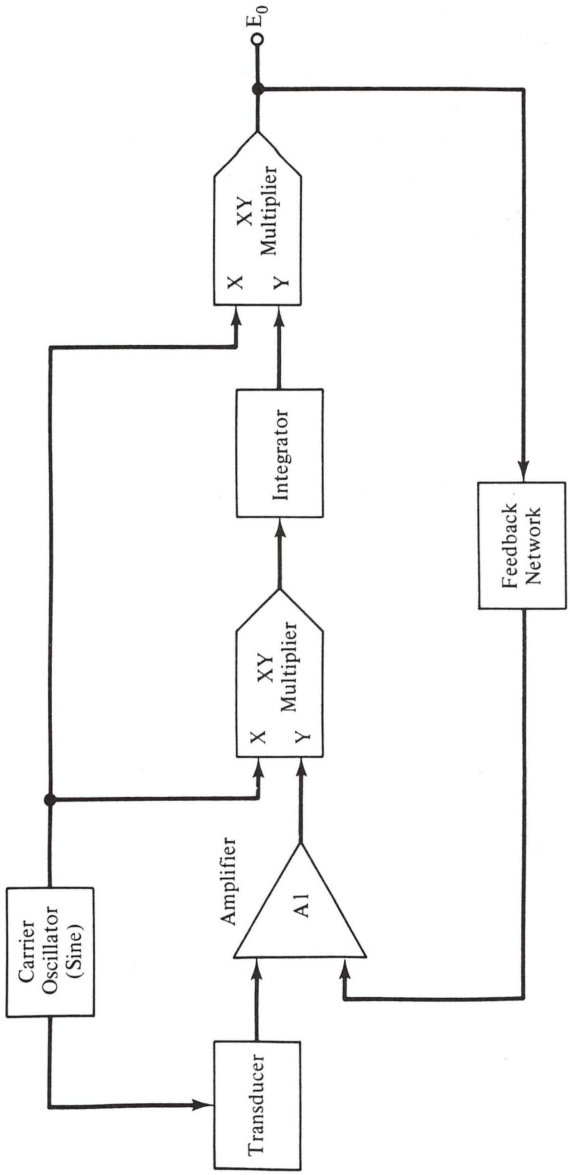


Figure 11-8. Autocorrelation amplifier.

chapter 12

OPERATIONAL AND LABORATORY AMPLIFIERS

12-1 OBJECTIVES

1. To learn basic operational amplifier theory.
2. To learn the major operational amplifier configurations used in instrumentation circuits.
3. To learn to derive the elementary operational amplifier transfer functions.
4. To learn how to use operational amplifiers in instrumentation.

12-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Write the transfer equation for (a) *inverting follower,* (b) *noninverting* follower.
2. Draw a circuit for an active *integrator* using an operational amplifier.
3. How would you cure *offset voltage* problems?
4. Write the transfer equation for an active operational amplifier *differentiator*.
5. What should be the value of a *compensation resistor,* R_c?

12-3 OPERATIONAL AMPLIFIERS: AN INTRODUCTION

The operational amplifier has been in existence for several decades, but only in the last ten or fifteen years has it come into its own as an almost universal electronic building block. The term *operational* is derived from the fact that these devices were originally designed for use in analog computers to solve *mathematical operations*. The range of circuit applications today, however, has increased immensely, so the operational amplifier has survived and prospered, even though analog computers, in which they were once a principal constituent, are now almost in eclipse.

Keep in mind, however, that even though the programmable analog computer is no longer used extensively, many instruments are little more than a nonprogrammable, dedicated-to-one-chore analog computer with a numeric readout of some sort.

In this chapter we will examine the *gross,* or large-scale, properties of the basic operational amplifier, and will learn to derive the transfer equations for most common operational amplifier circuits using only Ohm's law, Kirchoff's law, and the basic properties of all operational amplifiers.

One of the profound beauties of the modern, integrated-circuit, operational amplifier is its simplicity when viewed from the outside world. Of course, the inner workings are complex, but they are of little interest in our discussion of the operational amplifier's gross properties. We will limit our discussion somewhat by considering the operational amplifier as a *black box,* and that allows for a very simple analysis in which we relate the performance to the universal transfer function for all electronic circuits, namely, $E_{\text{out}}/E_{\text{in}}$.

12-4 PROPERTIES OF THE IDEAL OP-AMP

An *ideal* operational amplifier is a *gain block,* or black box if you prefer, that has the following general properties:

1. Infinite *open-loop* (i.e., no feedback) gain ($A_{\text{vol}} = \infty$)
2. Infinite input impedance ($Z_{\text{in}} = \infty$)
3. Zero output impedance ($Z_0 = 0$)
4. Infinite bandwidth ($f_0 = \infty$)
5. Zero noise generation

Of course, it is not possible to obtain a real IC operational amplifier that meets these properties—they are *ideal*—but if we read "infinite" as "very, very high," and "zero" as "very, very low," then the approximations of the ideal situation are very accurate. Real IC operational amplifiers, for example, can have an open-loop voltage gain from 50,000 to over 1,000,000, so it can be classed as *relatively infinite,* so the equations work in most cases.

12-5 DIFFERENTIAL INPUTS

Figure 12-1 shows the basic symbol for the common operational amplifier, including power terminals. In many schematics of operational amplifier circuits,

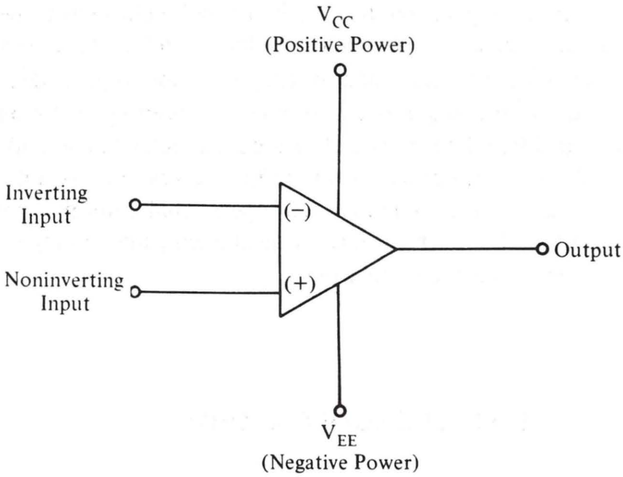


Figure 12-1. Symbol for an operational amplifier.

the V_{cc} and V_{ee} power terminals are deleted, so the drawing will be less "busy."

Note that there are two input terminals, labeled (−) and (+). The terminal labeled (−) is the *inverting* input. The output signal will be out of phase with signals applied to this input terminal; i.e., there will be a 180-degree phase shift. The terminal labeled (+) is the *noninverting* input, so output signals will be in phase with signals applied to this input. It is important to remember that these inputs look into *equal* open-loop gains, so they will have equal but opposite effects on the output voltage.

At this point let us add one further property to our list of ideal properties:

6. Differential inputs follow each other.

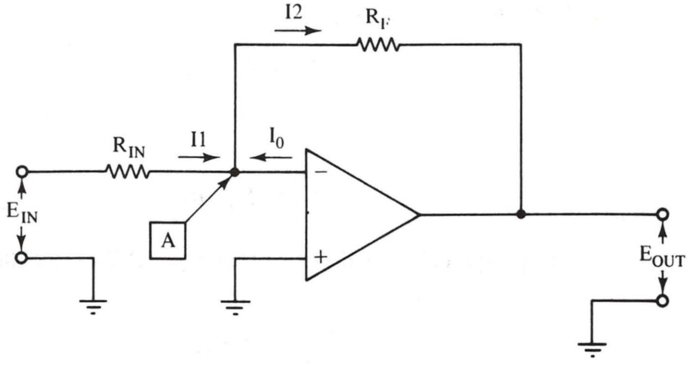


Figure 12-2. Inverting follower.

This property implies that the two inputs will behave as if they were at the same potential, especially under static conditions. In Fig. 12-2 we see an inverting follower circuit in which the noninverting (+) input is grounded. The sixth property allows us, in fact requires us, to treat the inverting (−) input as if it were *also* grounded. Many textbooks and magazine articles like to call this phenomenon a "virtual" ground, but that is a term that serves only to confuse the reader. It is better to accept as a basic axiom of operational amplifier circuitry that, for purposes of calculation and voltage measurement, the (−) input will be grounded if the (+) input is actually grounded.

12-6 ANALYSIS USING KIRCHOFF AND OHM

We know from Kirchoff's current law that the algebraic sum of all currents entering and leaving a point in a circuit must be zero. The total current flow into and out of point A in Fig. 12-2, then, must be *zero*. Three possible currents exist at this point: input current $I1$, feedback current $I2$, and any currents flowing into or out of the (−) input terminal of the operational amplifier, I_0. But according to ideal property 2, the input impedance of this type of device is infinite. Ohm's law tells us that by

$$I_0 = E/Z_{in} \tag{12-1}$$

current I_0 is zero, because E/Z_{in} is zero. So, if current I_0 is equal to zero, we conclude that $I1 + I2 = 0$ (Kirchoff's law). Since this is true, then

$$I2 = -I1 \tag{12-2}$$

We also know that

$$I1 = E_{in}/R_{in} \tag{12-3}$$

and

$$I2 = E_{out}/R_f \tag{12-4}$$

By substituting Eqs. (12-3) and (12-4) into Eq. (12-2), we obtain the result

$$\frac{E_{out}}{R_f} = \frac{-E_{in}}{R_{in}} \tag{12-5}$$

Solving for E_{out} gives us the transfer function normally given in operational amplifier literature for an inverting amplifier:

$$E_{\text{out}} = -E_{\text{in}} \times \frac{R_{\text{f}}}{R_{\text{in}}} \tag{12-6}$$

Example 12-1

Calculate the output voltage from an inverting operational amplifier circuit if the input signal is 100 mV, the feedback resistor is 100 kΩ, and the input resistor is 10 kΩ.

SOLUTION

$$\begin{aligned} E_{\text{out}} &= E_{\text{in}} \times (R_{\text{f}}/R_{\text{in}}) && \text{(12-6)} \\ E_{\text{out}} &= (0.1\ \text{V})(10^5\ \Omega/10^4\ \Omega) \\ E_{\text{out}} &= (0.1\ \text{V})(10) = \mathbf{1\ V} \end{aligned}$$

The term $R_{\text{f}}/R_{\text{in}}$ is the voltage gain factor, and is usually designated by the symbol A_{v}, which is written as

$$A_{\text{v}} = -R_{\text{f}}/R_{\text{in}} \tag{12-7}$$

We sometimes encounter Eq. (12-6) written using the left-hand side of Eq. (12-7):

$$E_{\text{out}} = A_{\text{v}}E_{\text{in}} \tag{12-8}$$

When designing simple inverting followers using operational amplifiers, use Eqs. (12-7) and (12-8). Let us look at a specific example. Suppose that we have a requirement for an amplifier with a gain of fifty. We want to drive this amplifier from a source that has an output impedance of 1000 ohms. A standard rule of thumb for designers to follow is to make the input impedance not less than ten times the source impedance, so in this case the amplifier must have an input source impedance that is equal to or greater than 10,000 ohms (10 kΩ). This requirement sets the value of the input resistor at 10 kΩ or higher, but in this example we select a 10 kΩ value for R_{in}.

$$\begin{aligned} A_{\text{v}} &= R_{\text{f}}/R_{\text{in}} && \text{(12-9)} \\ 50 &= R_{\text{f}}/10{,}000\ \text{ohms} && \text{(12-10)} \\ R_{\text{f}} &= 500{,}000\ \text{ohms} \end{aligned}$$

Our gain-of-fifty amplifiers will look like Fig. 12-3.

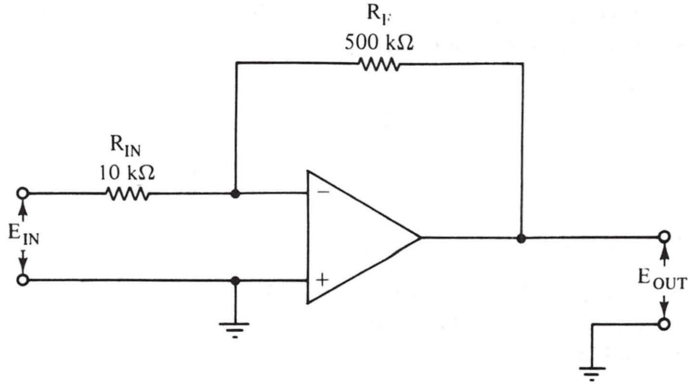


Figure 12-3. Gain-of-50 inverting follower.

12-7 NONINVERTING FOLLOWERS

The inverting follower circuits of Figs. 12-2 and 12-3 suffer badly from low input impedance, especially at higher gains, because the input impedance is the value of R_{in}. This problem becomes especially acute when we attempt to obtain even moderately high gain figures from low-cost devices. Although some types of operational amplifier allow the use of 500-kΩ to 2-megohm input resistors, they are costly and often uneconomical. The *noninverting follower* of Fig. 12-4 solves this problem by using the input impedance problem very nicely, because the input impedance of the op-amp is typically very, very high (ideal property 2).

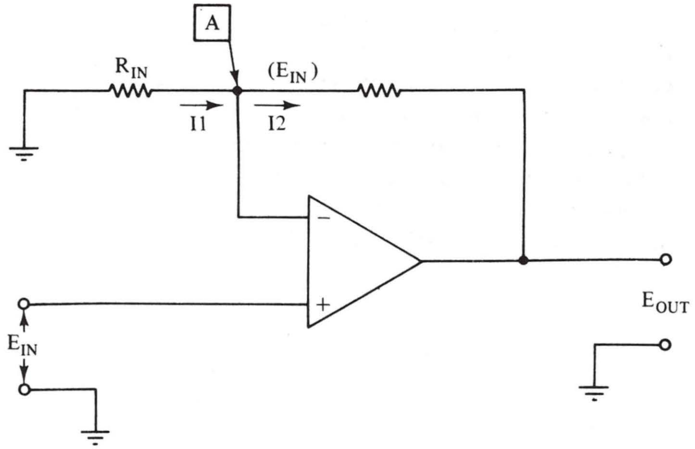


Figure 12-4. Noninverting follower with gain.

We may once again resort to Kirchoff's law to derive the transfer equation from our basic ideal properties. By property 6 we know that the inputs tend to follow each other, so the inverting input can be treated as if it were at the same potential as the noninverting input, which is E_{in}, the input signal voltage. We know that

$$I1 = I2 \tag{12-11}$$

$$I1 = E_{in}/R_{in} \tag{12-12}$$

$$I2 = (E_{out} - E_{in})/R_f \tag{12-13}$$

By substituting Eqs. (12-12) and (12-13) into Eq. (12-11) we obtain

$$I1 = I2 \tag{12-14}$$

$$\frac{E_{in}}{R_{in}} = \frac{E_{out} - E_{in}}{R_f} \tag{12-15}$$

Solving Eq. (12-15) for E_{out} results in the transfer equation for the noninverting follower amplifier circuit.

Multiply both sides by R_f

$$\frac{R_f E_{in}}{R_{in}} = E_{out} - E_{in} \tag{12-16}$$

Add E_{in} to both sides

$$\frac{R_f E_{in}}{R_{in}} + E_{in} = E_{out} \tag{12-17}$$

Factor out E_{in}

$$E_{in} \times \left[\frac{R_f}{R_{in}} + 1\right] = E_{out} \tag{12-18}$$

Example 12-2

Calculate the output voltage for 100 mV (i.e., 0.1 V) input in a noninverting follower amplifier if R_f is 100 kΩ and R_{in} is 10 kΩ.

SOLUTION

$$E_{out} = E_{in}\left[(R_f/R_{in}) + 1\right] \tag{12-18}$$

$$E_{out} = (0.1\ \text{V})\left[(10^5\ \Omega/10^4\ \Omega) + 1\right]$$

$$E_{out} = (0.1\ \text{V})(10 + 1)$$

$$E_{out} = (0.1\ \text{V})(11) = \mathbf{1.1\ V}$$

In this discussion we have arrived at both of the transfer functions commonly used in operational amplifier design, using only the basic properties, Ohm's law, and Kirchoff's current law. We may safely assume that the operational amplifier is merely a feedback device that generates a current that exactly cancels the input current. Figure 12-5 gives a synopsis of the characteristics of the most popular operational amplifier configurations. The unity gain noninverting follower of Fig. 12-5(c) is a special case of the circuit in Fig. 12-5(b), in which $R_f/R_{in} = 0$. In this case, the transfer equation becomes

$$E_{out} = E_{in}(0 + 1) \tag{12-19}$$

$$E_{out} = E_{in}(1) \tag{12-20}$$

$$E_{out} = E_{in} \tag{12-21}$$

12-8 OPERATIONAL AMPLIFIER POWER SUPPLIES

Although almost every circuit using operational amplifiers uses a dual polarity power supply, it is possible to operate the device with a single polarity supply. An example of single supply operation might be in equipment designed for mobile operation, or in circuits where the other circuitry requires only a single polarity supply, and an op-amp or two are but minority features in the design. It is, however, generally better to use the bipolar supplies as intended by the manufacturer.

There are two separate power terminals on the typical operational amplifier device, and these are marked V_{cc} and V_{ee}. The V_{cc} supply is connected to a power supply that is *positive* to ground, while the V_{ee} supply is *negative* with respect to ground. These supplies are shown in Fig. 12-6. Keep in mind that, although batteries are shown in the example, regular power supplies may be used instead. Typical values for V_{cc} and V_{ee} range from ± 3 volts dc to ± 22 volts dc. In many cases, perhaps most, the value selected for these potentials will be between ± 9 volts dc and ± 15 volts dc.

There is one further constraint placed on the operational amplifier power supply: $V_{cc} - V_{ee}$ must be less than some specified voltage, usually 30 volts. So, if V_{cc} is $+18$ volts dc, then V_{ee} must be not greater than $(30 - 18)$, or 12 volts dc.

12-9 PRACTICAL DEVICES: SOME PROBLEMS

Before we can properly apply operational amplifiers in real equipment we must learn some of the limitations of real-world devices. The devices that we have considered up until now have been *ideal,* so they do not exist. Real IC operational amplifiers carry price tags of less than half a dollar up to several dozen dollars each. The lower the cost, generally, the less ideal the device.

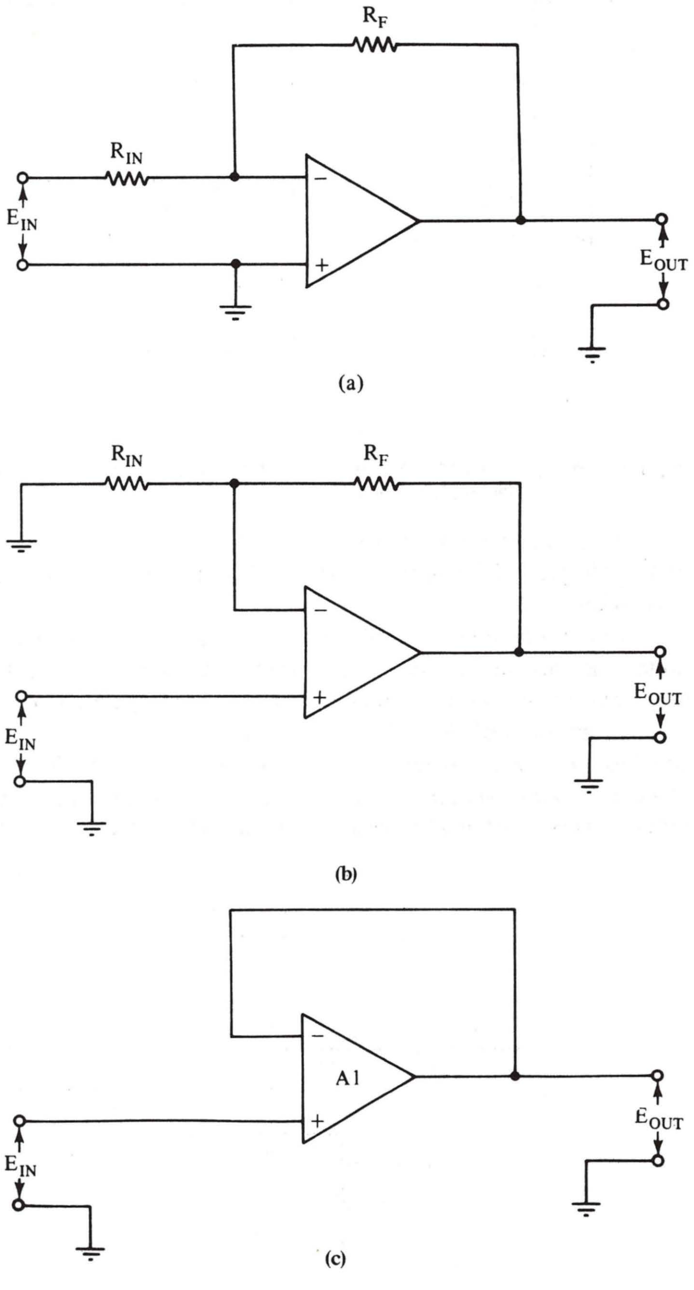


Figure 12-5. (a) Inverting follower. (b) Noninverting gain follower. (c) Unity gain noninverting follower.

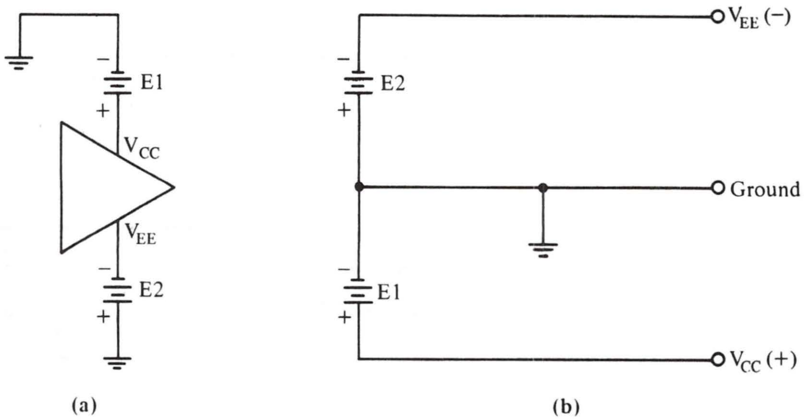


Figure 12-6. (a) Power supply requirements of an operational amplifier. (b) Typical operational amplifier configuration.

Three main problems exist in real operational amplifiers: offset current, offset voltage, and frequency response. Of less importance in many cases is noise generation.

In real operational amplifier devices the input impedance is less than infinite, and this implies that a small input bias current exists. The input current may flow into or out of the input terminals of the operational amplifier. In other words, current I_0 of Fig. 12-2 is *not* zero, so it will produce an output voltage equal to $-I_0 \times R_f$. The cure for this problem is shown in Fig. 12-7, and involves placing a *compensation resistor* between the noninverting input terminal and ground. This tactic works because the currents in the respective inputs are ap-

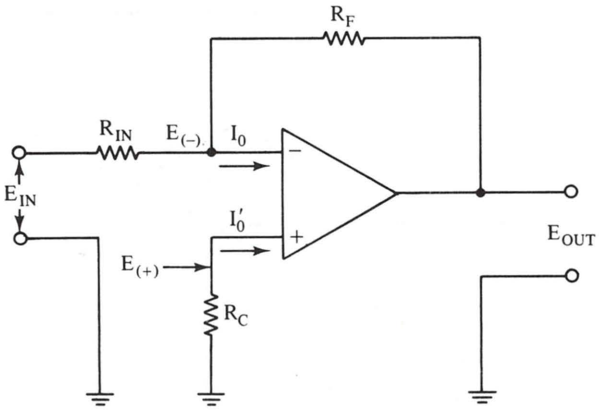


Figure 12-7. Use of a compensation resistor.

proximately equal. Since resistor R_c is equal to the parallel combination of R_f and R_{in}, it will generate the same *voltage drop* that appears at the inverting input. The resultant output voltage, then, is zero, because the two inputs have equal but opposite polarity effect on the output.

Output offset voltage is the value of E_{out} that will exist if the input end of R_{in} is grounded (i.e., $E_{in} = 0$). In the ideal device, E_{out} would be zero under this condition, but in real devices there may be some offset potential present. This output potential can be forced to zero by any of the circuits in Fig. 12-8.

The circuit in Fig. 12-8(a) uses a pair of *offset null* terminals found on many, but not all, operational amplifiers. Although many IC operational amplifiers use this technique, some do not. Alternatively, the offset range may be insufficient in some cases. In either event, we may use the circuit of Fig. 12-8(b) to solve the problem.

The offset null circuit of Fig. 12-8(b) creates a current flowing in resistor $R1$ to the summing junction of the operational amplifier. Since the offset current may flow either *into* or *out of* the input terminal, the null control circuit must be able to supply currents of both polarities. Because of this requirement, the ends of the potentiometer ($R1$) are connected to V_{cc} and V_{ee}.

In many cases, it is found that the offset is small compared with normally expected values of input signal voltage. This is especially true in low-gain applications, in which case the nominal offset current will create such a low output error that no action need be taken. In still other cases, the offset of each stage in a cascade chain of amplifiers may be small, but their cumulative effect may be a large offset error. In this type of situation, it is usually sufficient to null only one of the stages late in the chain (i.e., close to the output stage).

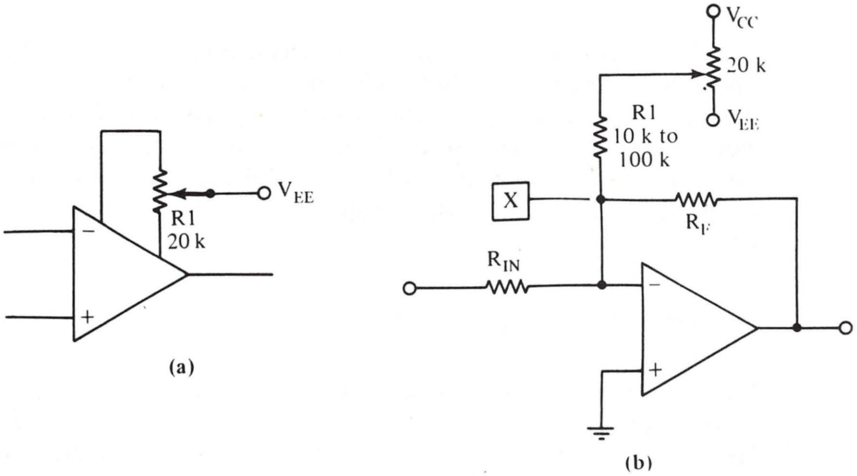
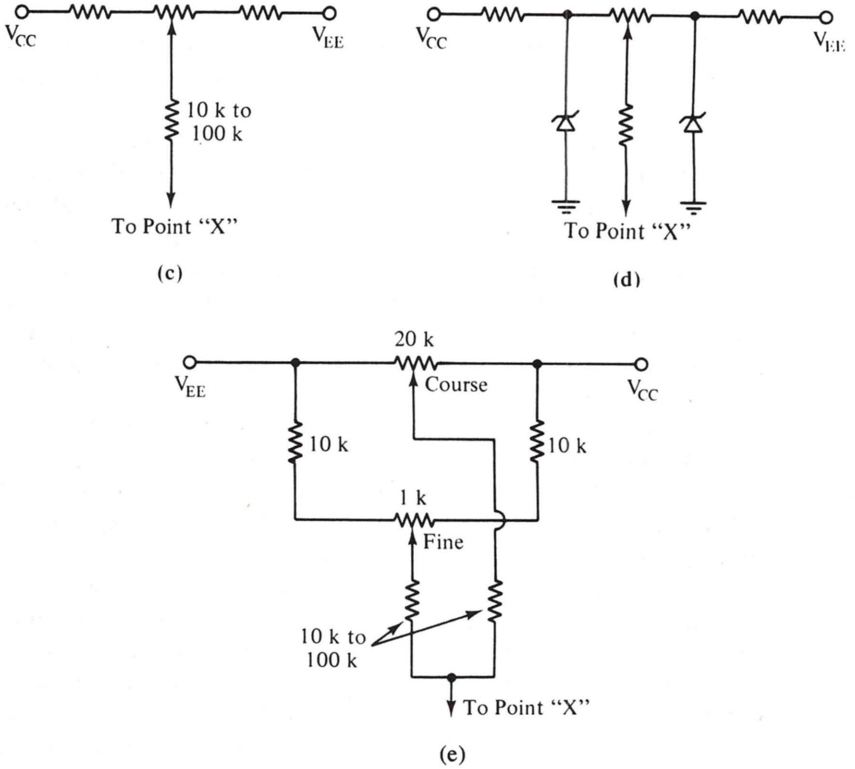


Figure 12-8. (a) Use of offset terminals to null output. (b) Use of summing current to null output.



(c)

(d)

(e)

Figure 12-8. *(cont.)* (c)–(e) High resolution offset null circuits.

In those circuits where the offset is small, but critical, it may be useful to replace $R1$ and $R2$ of Fig. 12-8(b) with one of the resistor networks of Figs. 12-8(c) through 12-8(e). These perform essentially the same function, but have superior resolution. That is to say, there is a smaller change in output voltage for a single turn of the potentiometer. This type of circuit will have a superior resolution in any event, but even further improvement is possible if a ten-turn (or more) potentiometer is used.

12-10 DC DIFFERENTIAL AMPLIFIERS

The fact that an IC operational amplifier has two complementary inputs, inverting and noninverting, makes it a natural for application as a *differential amplifier*. These circuits produce an output voltage that is proportional to the *difference* between two ground-referenced input voltages. Recall from our previous discussion that the two inputs of an operational amplifier have equal but opposite effect on the output voltage. If the same voltage or two equal voltages are applied to the two inputs (i.e., a *common mode* voltage, $E3$ in Fig. 12-1), then the output voltage will be zero. The transfer equation for a differential amplifier is

$$E_{out} = A_v(E1 - E2) \tag{12-22}$$

So, if $E1 = E2$, then $E_{out} = 0$.

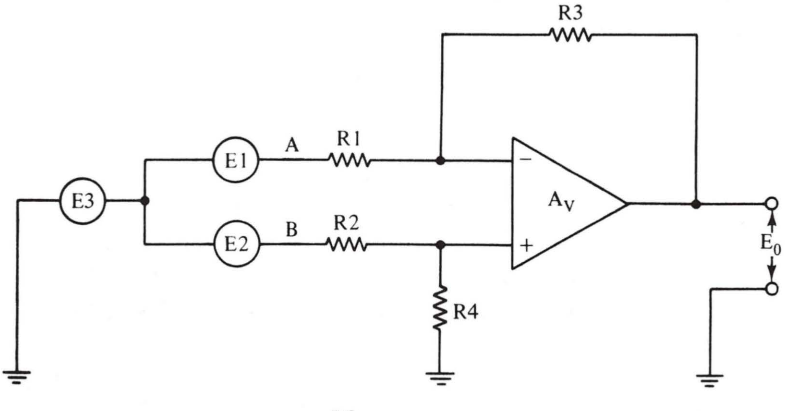
The circuit of Fig. 12-9 shows a simple differential amplifier using a single IC operational amplifier. The voltage gain of this circuit is given by

$$A_v = R3/R1 \tag{12-23}$$

provided that $R1 = R2$ and $R3 = R4$.

The main appeal of this circuit is that it is economical, requiring but one IC operational amplifier. It will reject common mode voltages reasonably well if the equal resistors are well matched. A glaring problem exists, however, and that is a low input impedance. Additionally, with the problems existing in real operational amplifiers, this circuit may be a little difficult to tame in high-gain applications. As a result, designers frequently use an alternate circuit in these cases.

In recent years, the instrumentation amplifier (I.A.) of Fig. 12-10 has become popular because it alleviates most of the problems associated with the circuit of Fig. 12-9. The input stages are noninverting followers, so they will have a characteristically high input impedance. Typical values run to as much as 1000 megohms.



$$A_V = \frac{R3}{R1}$$

Provided: R1 = R2
R3 = R4

Figure 12-9. Differential amplifier with input voltages.

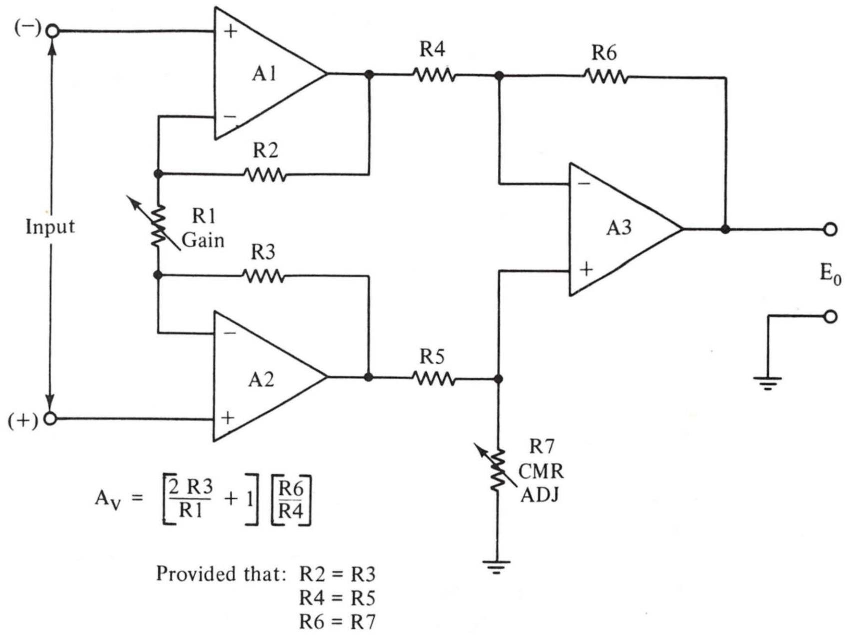


Figure 12-10. Instrumentation amplifier.

The instrumentation amplifier is relatively tolerant of different resistor ratios used to create voltage gain. In the simplest case, the differential voltage gain is given by

$$A_v = \frac{2R3}{R1} + 1 \tag{12-24}$$

provided that $R3 = R2$, and $R4 = R5 = R6 = R7$.

It is interesting to note that the common-mode rejection ratio is not seriously degraded by mismatch of resistors $R2$ and $R3$; only the gain is affected. If these resistors are mismatched, then a differential voltage gain error will be introduced.

The situation created by Eq. (12-24) results in having the gain of $A3$ equal to unity (i.e., 1), and that is a waste. If gain in $A3$ is desired, then Eq. (12-24) must be rewritten into the form

$$A_v = \left[\frac{2R3}{R1} + 1\right]\left[\frac{R7}{R5}\right] \tag{12-25}$$

Example 12-3

Calculate the differential voltage gain of an instrumentation amplifier that uses the following resistor values: $R3 = 33\ \text{k}\Omega$, $R1 = 2.2\ \text{k}\Omega$, $R5 = 3.3\ \text{k}\Omega$, and $R7 = 15\ \text{k}\Omega$.

SOLUTION

$$A_v = [(2R3/R1) + 1)(R7/R5)] \tag{12-25}$$

$$A_v = \left[\frac{(2)(33\ \text{k}\Omega)}{(2.2\ \text{k}\Omega)} + 1\right]\left[\frac{(15\ \text{k}\Omega)}{(3.3\ \text{k}\Omega)}\right]$$

$$A_v = \mathbf{141}$$

One further equation that may be of interest is the general expression from which the other instrumentation amplifier transfer equations are derived:

$$A_v = \frac{R7(R1 + R2 + R3)}{R1R6} \tag{12-26}$$

which remains valid provided that the ratio $R7/R6 = R5/R4$.

Equation (12-26) is especially nice, since you need not be concerned with matched pairs of precision resistors, but only that their ratios be equal.

12-11 PRACTICAL CIRCUIT

In this section we will consider a practical design example using the instrumentation amplifier circuit. The particular problem required a frequency response to 100 kHz, and that the input lines be shielded. But the latter requirement would also deteriorate the signal at high frequencies because of the shunt capacitance of the input cables. To overcome this problem a *high-frequency compensation* control is built into the amplifier. Voltage gain is approximately ten.

The circuit to the preamplifier is shown in Fig. 12-11. It is, of course, the instrumentation amplifier of Fig. 12-10 with some modifications. When the frequency response is less than 10 kHz or so, we may use any of the 741-family devices (i.e., 741, 747, 1456, and 1458), but premium performance demands a better operational amplifier. In this case, one of the most economical is the RCA CA3140, although an L156 would also suffice.

Common mode rejection can be adjusted to compensate for any mismatch in the resistors or IC devices by adjusting $R10$. This potentiometer is adjusted for zero output when the same signal is applied simultaneously to both inputs.

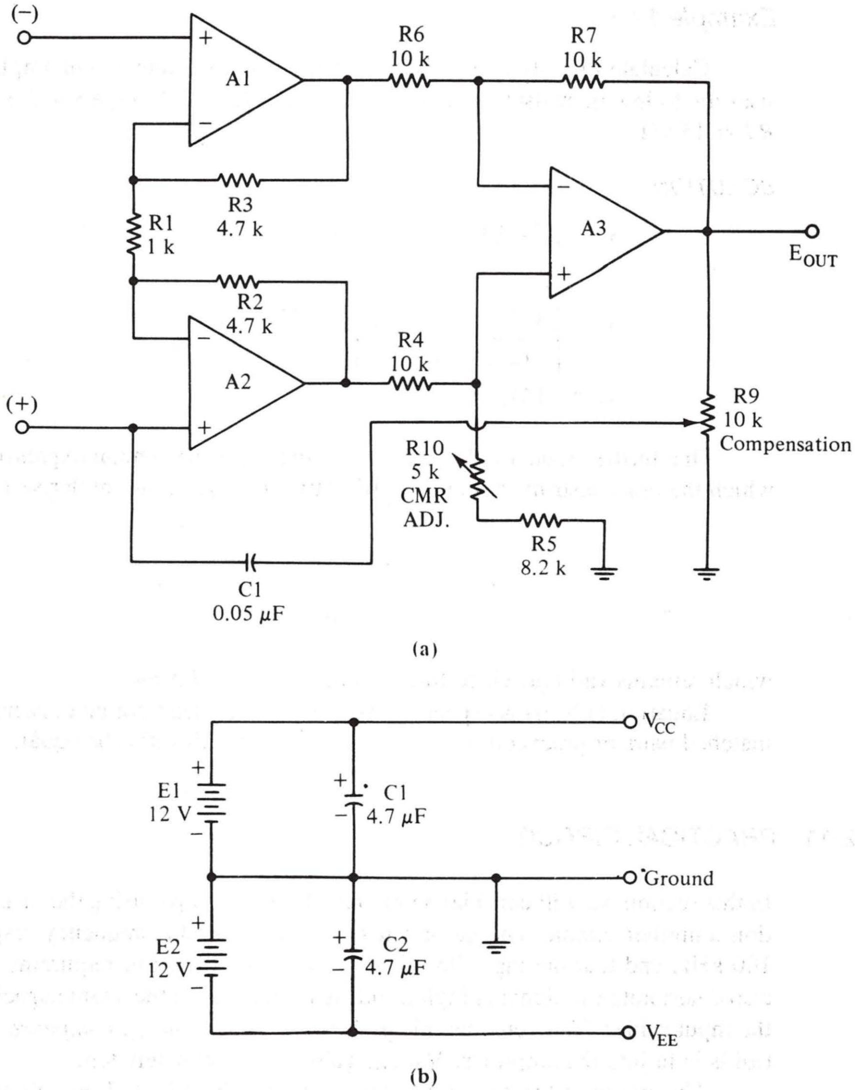


Figure 12-11. (a) Instrumentation amplifier with capacitance null. (b) Power supply for (a).

The frequency response characteristics of this preamplifier are shown in Figs. 12-12 through 12-16. The input in each case was a 1000-Hz square wave from a function generator. The waveform in Fig. 12-12 shows the output signal when resistor *R*9 is set with its wiper closest to ground. Notice that it is essentially square, showing only a small amount of roll-off of high frequencies. The

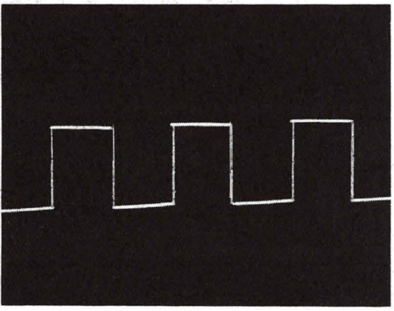


Figure 12-12. Square wave.

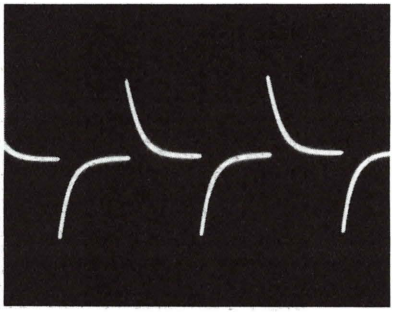


Figure 12-13. Differentiated square wave.

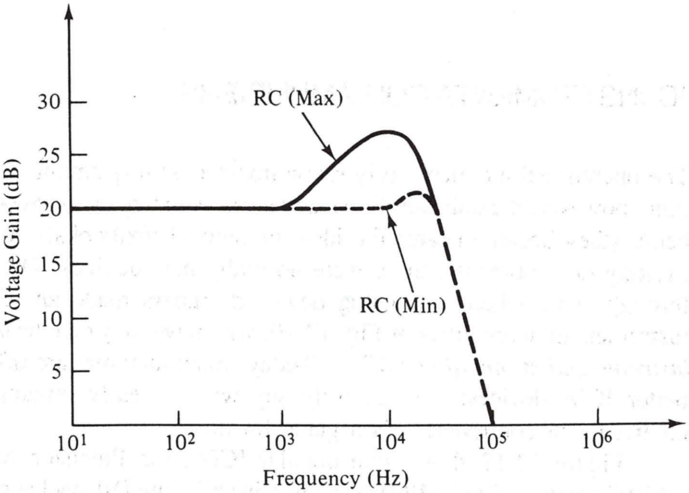


Figure 12-14. Voltage gain-vs-frequency.

waveform in Fig. 12-13 is the same signal when $R9$ is at maximum resistance. This creates a small amount of regenerative (i.e., positive) feedback, although it is not sufficient to start oscillation, but will enhance amplification of high frequencies.

The problem of oscillation can be quite serious (Fig. 12-15), however, if certain precautions are not taken, most of which involve limiting the amplitude of the feedback signal. This goal is realized by using a 2200-ohm resistor in series with the potentiometer.

Another source of oscillation is the value of $C1$. When a 0.001-μF capacitor is used at $C1$, an 80-kHz oscillation is created (see Fig. 12-16). The frequency response is shown in Fig. 12-14. To obtain any particular response curve, modify the values of $C1$ and $R9$.

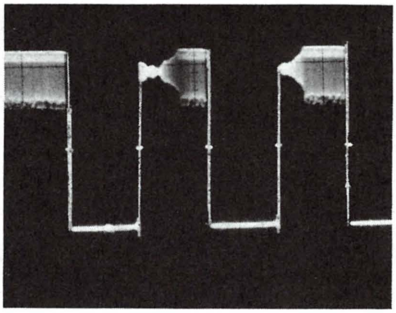


Figure 12-15. Ringing on square wave.

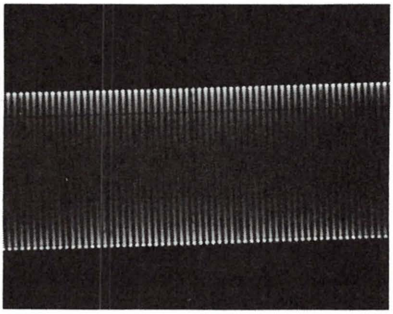


Figure 12-16. Eighty-kHz oscillation.

12-12 IC INSTRUMENTATION AMPLIFIERS

The operational amplifier truly revolutionized analog circuit design. For a long time, however, the only additional advances were that op-amps became better and better (they became nearer the ideal op-amp of textbooks!). While those were exciting developments, there were no really new devices. The next big breakthrough came when the analog device designers made an IC version of the instrumentation amplifier of Fig. 12-10, which we may call the *integrated circuit instrumentation amplifier* (ICIA). Today, manufacturers are offering better and better ICIA devices; we can truly say with an early op-amp textbook "the contriving of contrivances is a game for all."

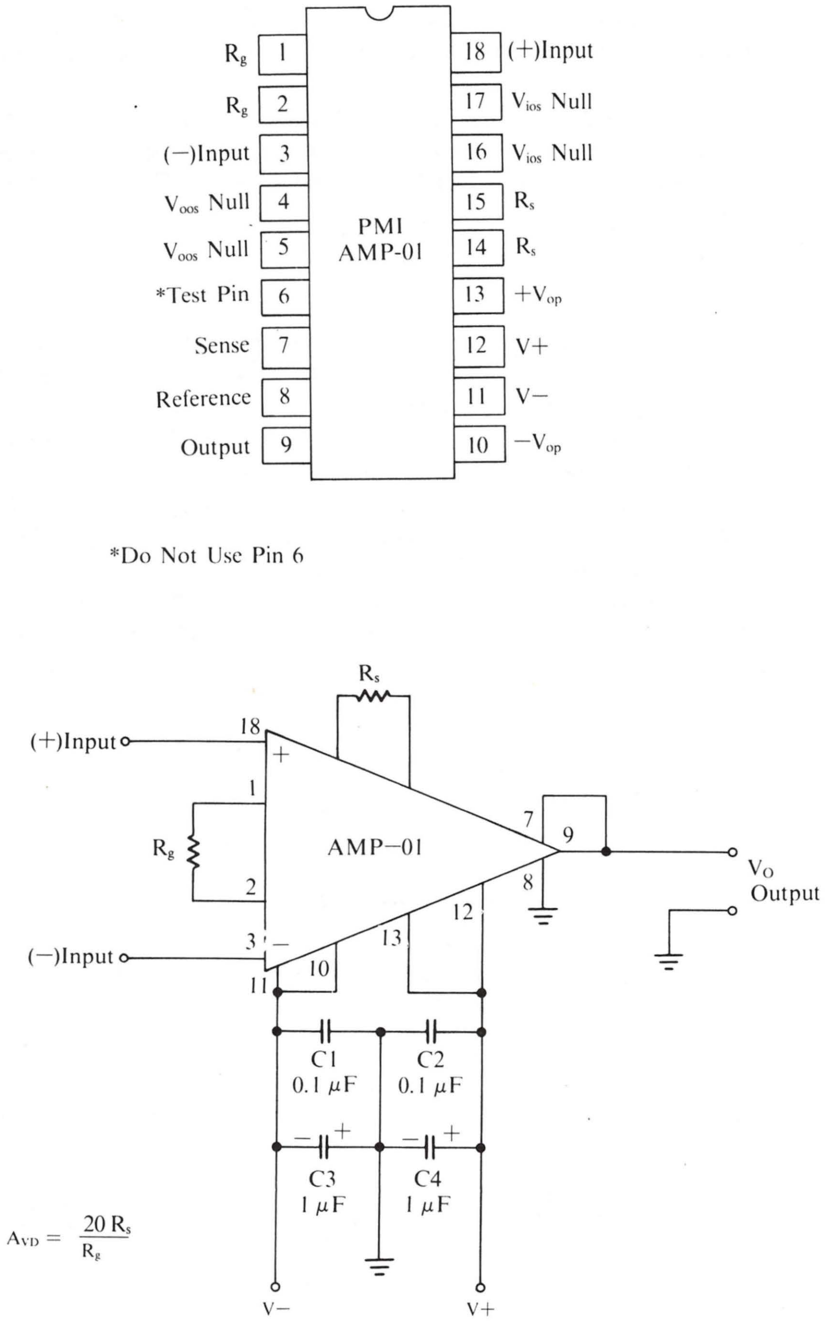
Figure 12-17 shows one popular ICIA, the Precision Monolithics, Inc. AMP-01 device. The AMP-01 is housed in an 18-pin DIP package (Fig. 12-17(a)).

The basic circuit for the AMP-01 is shown in Figure 12-17(b). Notice how simple the circuit is! There are few connections: differential inputs, dc power supplies (V− and V+), output, ground, and two gain-setting resistors. The voltage gain of this circuit is given by:

$$A_{vd} = 20R_s/R_g \qquad (12\text{-}27)$$

Suppose we want to make a differential voltage amplifier with a gain of ×1000. We need to make a resistor ratio of 1000/20, or 50:1. Thus, if R_s is set to 100 kohms and R_g is 2 kohms, we will have the required gain of 1000. The permissible gain range is 0.1 to 10,000.

The dc power supply voltages are up to −/+ 18 volts dc. Notice in Fig. 12-17(b) that the dc power supply lines are heavily bypassed. The 0.1-μF units are used to bypass high frequencies, while the 1-μF units are for low frequencies. The 0.1-μF units must be mounted as close as possible to the body of the amplifier.



*Do Not Use Pin 6

$$A_{VD} = \frac{20\,R_s}{R_g}$$

Figure 12-17. Precision Monolithics, Inc., AMP-01 ICIA: (a) pinouts, (b) circuit.

The maximum operating frequency depends upon the gain. At a gain of 1, the maximum small-signal input frequency is 570 kHz, while at a gain of 1000 it reduces to 26 kHz.

The Burr-Brown INA-101 is another ICIA device. This amplifier is also simple to connect. There are only dc power connections, differential input connections, offset adjust connections, ground, and an output. The gain of the circuit is set by:

$$A_{vd} = (40\ k/R_g) + 1 \tag{12-28}$$

The INA-101 is basically a low-noise, low input bias current integrated circuit version of the IA of Fig. 12-17(a). The resistors labeled $R2$ and $R3$ in Fig. 12-17(a) are 20 kohms, hence the "40 k" term in Fig. 12-18.

Potentiometer $R1$ in Fig. 12-18 is used to null the offset voltages appearing at the output. An *offset voltage* is a voltage that exists on the output at a time when it should be zero (i.e., when $V1 = V2$, so that $V1 - V2 = 0$). The offset voltage might be internal to the amplifier or a component of the input signal. Dc offsets in signals are common, especially in biopotential amplifiers such as ECG and EEG.

Still another ICIA is the LM-363 device shown in Fig. 12-19; the miniDIP version is shown in Fig. 12-19(a) (an 8-pin metal can is also available), while a

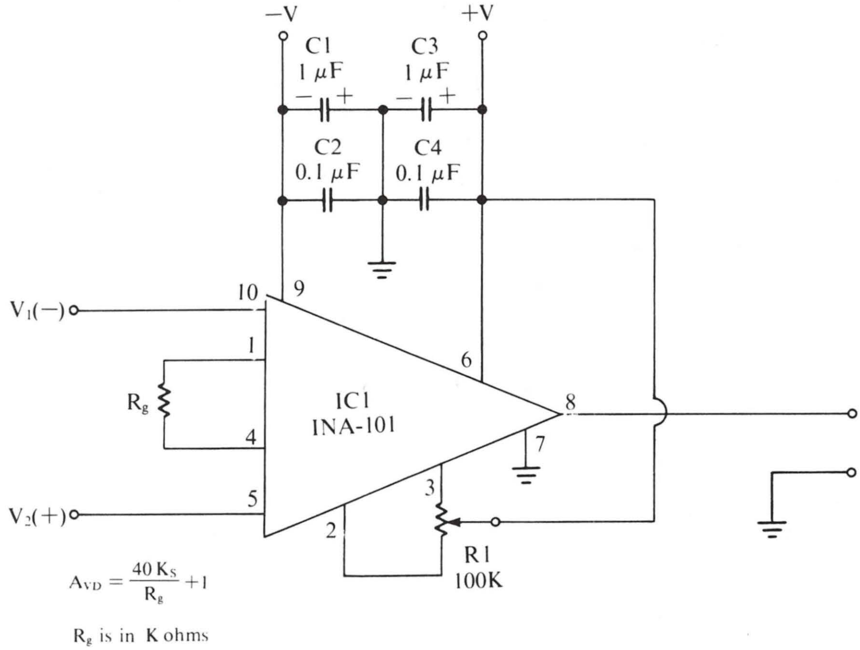


Figure 12-18. Burr-Brown INA-101 ICIA.

typical circuit is shown in Fig. 12-19(b). The LM-363 device is a fixed gain ICIA. There are three versions:

Designation	*Gain*
LM-363-10	10
LM-363-100	100
LM-363-500	500

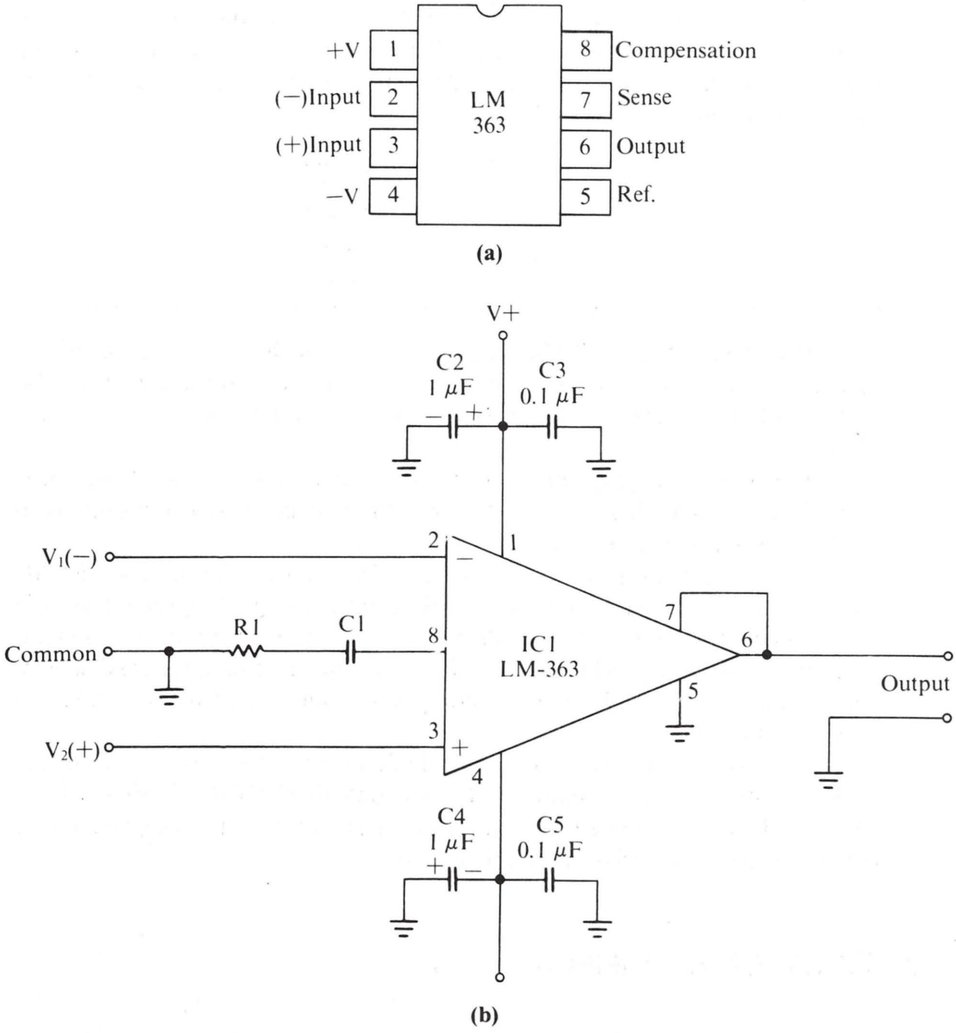


Figure 12-19. National Semiconductor LM-363 ICIA: (a) pinouts, (b) circuit.

The LM-363-x is useful in places where one of the standard gains is required and there is minimum of space available. Two examples spring to mind. We could use the LM-363-x as a transducer preamplifier, especially in noisy signal areas; the LM-363-x can be built onto (or into) the transducer to build up its signal before sending it to the main instrument or signal acquisition computer. The other example is in bioamplifiers. The biopotentials are typically very small, especially in lab animals. The LM-363-x can be mounted on the subject and a higher level signal can be sent to the main instrument; a little exotic, but nonetheless useful.

A selectable gain version of the LM-363 device is shown in Fig. 12-20(b); the 16-pin DIP package is shown in Fig. 12-20(a), while a typical circuit is shown in Fig. 12-29(b). The type number of this device is LM-363-AD, which distinguishes it from the LM-363-x devices. The gain can be ×10, ×100, or ×1000 depending upon the programming of the gain setting pins (2, 3, and 4). The programming protocol is as follows:

Gain Desired	*Jumper Pins*
× 10	(all open)
× 100	3 & 4
× 1000	2 & 4

Switch $S1$ in Fig. 12-20(b) is the *gain select* switch. This switch should be mounted close to the IC device, but is quite flexible in mechanical form. The switch could also be made from a combination of CMOS electronic switches (e.g., 4066).

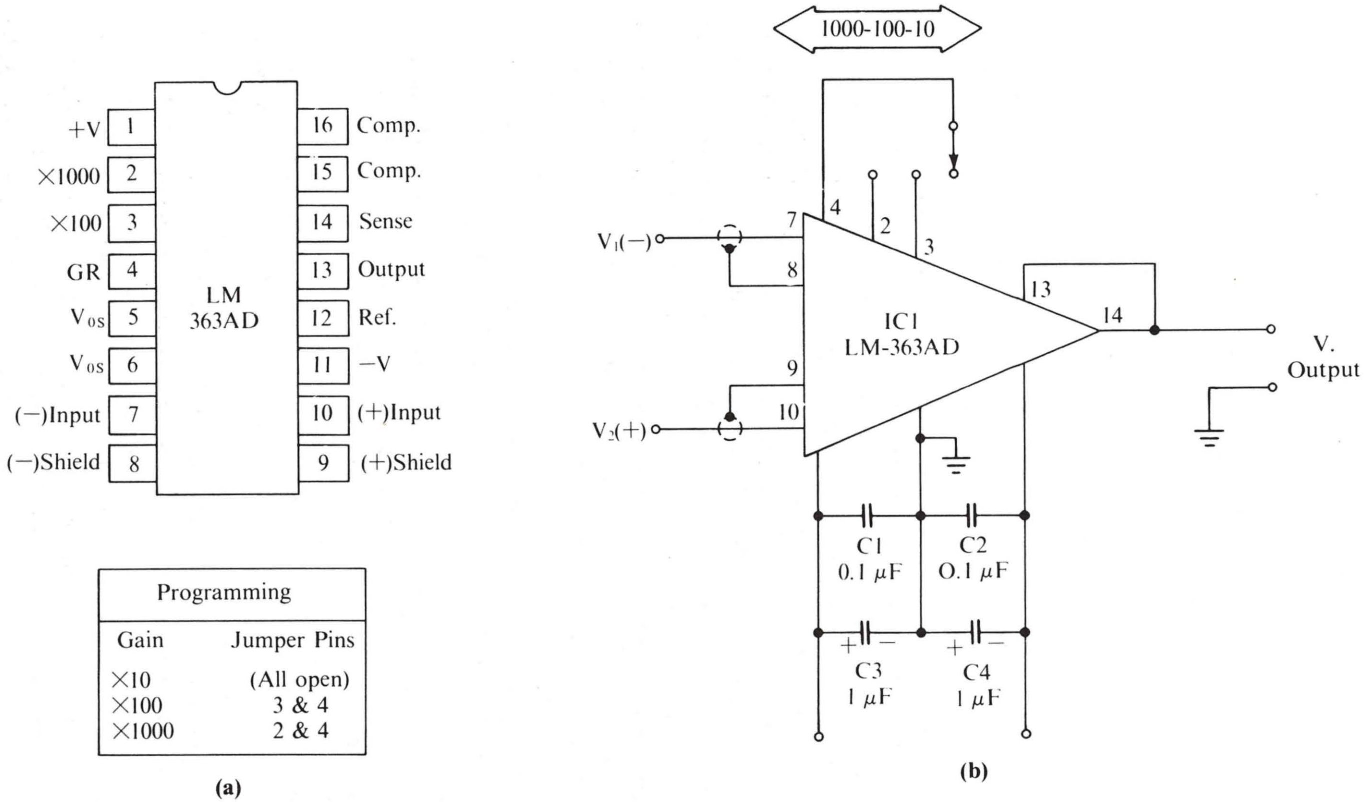
The dc power supply terminals are treated in a manner similar to the other amplifiers. Again, the 0.1-μF capacitors need to be mounted as close as possible to the body of the LM-363-AD.

Pins 8 and 9 are guard shield outputs. These pins make the LM-363-AD more useful for many instrumentation problems than other models. By outputting a signal sample back to the shield of the input lines, we can increase the common-mode rejection ratio. This feature is used a lot in bipotential amplifiers and in other applications where a low-level signal must pass through a strong interference (high noise) environment.

The LM-363 devices will operate with dc supply voltages of −/+ 5 volts to −/+ 18 volts dc, with a common-mode rejection ratio (CMRR) of 130 dB. The 7 nV/(SQR(Hz)) noise figure makes the device useful for low-noise applications (a 0.5 nV model is available at premium cost).

12-13 ISOLATION AMPLIFIERS

Many applications for instrumentation amplifiers are dangerous for either the circuit or the user. In biomedical applications, the issue is patient safety. There are



Programming	
Gain	Jumper Pins
×10	(All open)
×100	3 & 4
×1000	2 & 4

(a)

(b)

Figure 12-20. National Semiconductor LM-363AD ICIA: (a) pinouts, (b) circuit.

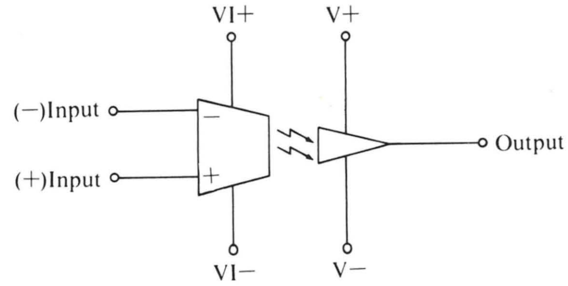


Figure 12-21. Isolation amplifier symbol.

numerous signal acquisition needs in biomedical instrumentation where the victim—errrr patient—is at risk. Even the simple ECG machine, which measures and records the heart's electrical activity, was once implicated in patient safety problems. Another problem area in biomedical applications is catheterization instruments. There are several tests where the doctors insert an electrode or transducer into the body and then measure the resulting signal: the intracardiac ECG places an electrode inside the heart by way of a blood vein; the cardiac output computer uses a signal from a thermistor inside a catheter placed in the heart (also through a vein); and simple electronic blood pressure monitors use a transducer that connects to an artery. In all of these cases, we do not want the patient exposed to small differences of potential due to current leakage from the 60 Hz ac power lines. The solution is use of an isolation amplifier.

Another application is signals acquisition in high voltage circuits. We do not want to mix high voltage sources with low voltage electronics because we don't want the low voltage circuits to blow out. Again, the solution is the isolation amplifier.

Figure 12-21 shows the basic symbol for the isolation amplifier. The break in the triangle used to represent any amplifier denotes the fact that there is an extremely high impedance (typically 10^{12} ohms) between the inputs and output terminal of the isolation amplifier.

Notice that there are two sets of dc power supply terminals. The V− and V+ terminals are the same as found on all ICIA or op-amp devices. These dc power supply terminals are connected to the regular dc supply of the equipment where the device is used. Such a power supply derives its dc potential from the ac power mains by way of a 60 Hz transformer. The isolated de power supply inputs (VI− and VI+) are used to power the input amplifier stages and must be isolated from the main dc power supply of the equipment. The VI− and VI+ terminals are usually either battery powered, or, powered from a dc-to-dc converter that produces a dc output from the main power supply by using a high frequency (50 to 500 kHz) oscillator. The high frequency "power supply" transformer does not pass 60 Hz signals well, so the isolation is maintained.

Figure 12-22 shows the circuit of an isolation amplifier based on the Burr-Brown 3652 device. This isolation amplifier is not generally available to hobbiests, but would be used even in small "one of a kind" professional labs.

The dc power for both the isolated and nonisolated sections of the 3652 is provided by the 722 dual dc-to-dc converter. This device produces two independent − / + 15 Vdc supplies that are each isolated from the 60 Hz ac power mains and from each other. The 722 device is powered from a + 12 Vdc source that is derived from the ac power mains. In some cases, the nonisolated section (which is connected to the output terminal) is powered from a bipolar dc power supply that is derived from the 60 Hz ac mains, such as a −/+ 12 Vdc or −/+ 15 Vdc supply. In no instance, however, should the isolated dc power supplies be derived from the ac power mains.

There are two separate ground systems in this circuit, symbolized by the small triangle and the regular three-bar "chassis" ground symbol. The isolated ground is not connected to either the dc power supply ground/common or the chassis ground. It is kept floating at all times and becomes the signal common for the input signal source.

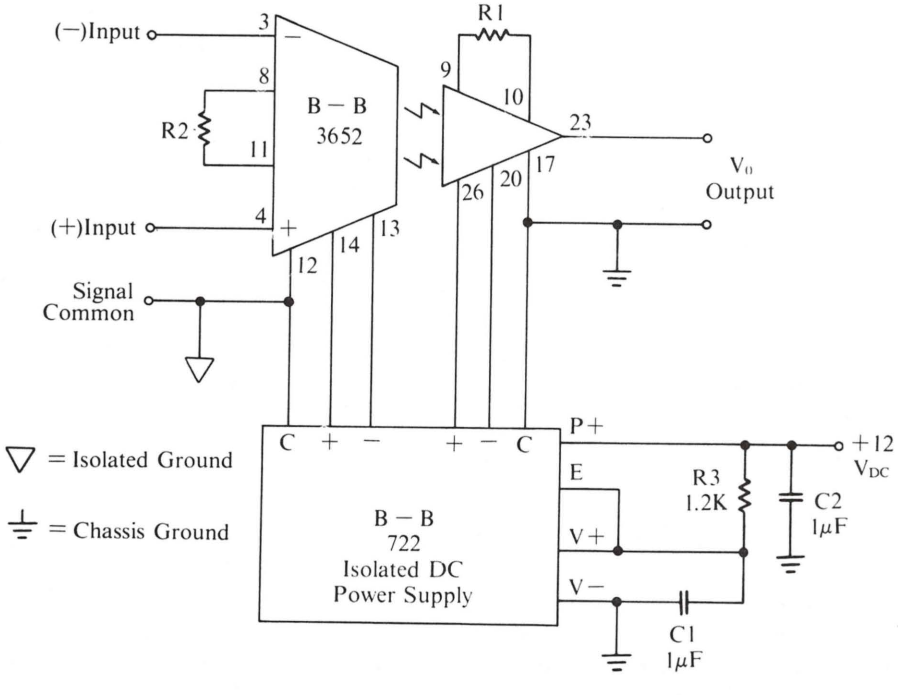


Figure 12-22. Isolation amplifier circuit.

The gain of the circuit is approximately:

$$\text{GAIN} = \frac{1{,}000{,}000}{R1 + R2 + 115} \tag{12-29}$$

In most design cases, the issue is the unknown values of the gain setting resistors. We can rearrange the equation above to solve for $(R1 + R2)$:

$$(R1 + R2) = \frac{1{,}000{,}000 - (115 \times \text{GAIN})}{\text{GAIN}} \tag{12-30}$$

where R1 and R2 are in ohms
GAIN is the voltage gain desired

Let's work an example. Suppose we need a differential voltage gain of 1,000. What combination of $R1$ and $R2$ will provide that gain figure? If GAIN = 1000

$$(R1 - R2) = \frac{1{,}000{,}000 - (115 \times 1000)}{1000} \tag{12-31}$$

$$(R1 + R2) = \frac{1{,}000{,}000 - (115{,}000)}{1000} \tag{12-32}$$

$$(R1 + R2) = \frac{885{,}000}{1000} \tag{12-33}$$

$$(R1 + R2) = 885 \text{ ohms} \tag{12-34}$$

In this case, we need some combination of $R1$ and $R2$ that adds to 885 ohms. The value 440 ohms is ''standard'' and will result in only a tiny gain error if used.

The IC instrumentation amplifier and the isolation amplifier open new applications that the simple op-amp cannot match.

12-14 DIFFERENTIAL AMPLIFIER APPLICATIONS

Differential amplifiers find application in many different instrumentation situations. Of course, it should be realized that they are required wherever a differential signal voltage is found. Less obvious, perhaps, is that they are used to acquire signals or to operate in control systems, in the presence of large noise signals. Many medical applications, for example, use the differential amplifier, because they look for minute biopotentials in the presence of strong 60-Hz fields from the ac power mains.

Another class of applications is the amplification of the output signal from a Wheatstone bridge, and this is shown in Fig. 12-23. If one side of the bridge's excitation potential is grounded, then the output voltage is a differential signal voltage. This signal can be applied to the inputs of a differential amplifier or instrumentation amplifier to create an amplified, single-ended, output voltage.

A "rear end" stage suitable for many operational amplifier instrumentation projects is shown in Fig. 12-24. This circuit consists of three low-cost operational amplifier ICs. Since they follow most of the circuit gain, we may use low-cost devices such as the 741 in this circuit. The gain of this circuit is given by $R2/10^4$.

12-15 INTEGRATORS

Figure 12-25 shows the basic operational amplifier *integrator* circuit. The transfer equation for this circuit may be derived in the same manner as before, with due consideration for $C1$.

$$I2 = -I1 \tag{12-35}$$

but

$$I1 = E_{\text{in}}/R1 \tag{12-36}$$

and

$$I2 = C1(dE_0/dt) \tag{12-37}$$

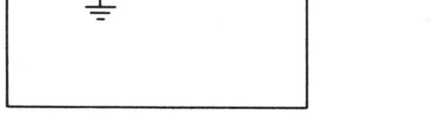


Figure 12-23. Differential amplifier used to amplify output of Wheatstone bridge.

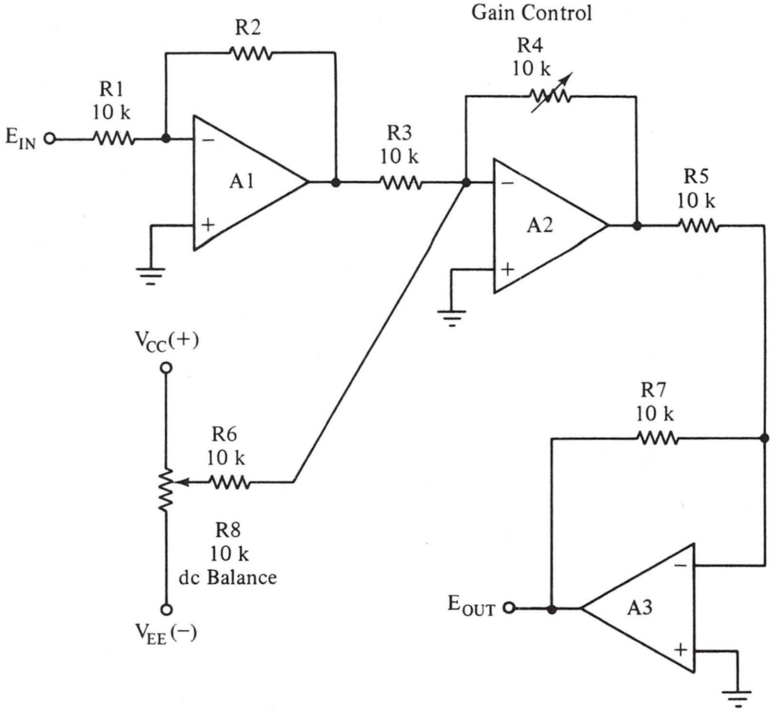


Figure 12-24. Universal "rear end" for instrumentation amplifiers and other purposes.

Substituting Eqs. (12-36) and (12-37) into Eq. (12-35) results in

$$\frac{C1\, dE_0}{dt} = \frac{-E_{\text{in}}}{R1} \tag{12-38}$$

We may now solve Eq. (12-38) for E_o by integrating both sides.

$$\int \frac{C1\, dE_0}{dt}\, dt = - \int \frac{E_{\text{in}}}{R1}\, dt \tag{12-39}$$

$$C1E_0 = \frac{-1}{R1} \int_0^t E_{\text{in}}\, dt \tag{12-40}$$

$$E_0 = \frac{-1}{R1C1} \int_0^t E_{\text{in}}\, dt \tag{12-41}$$

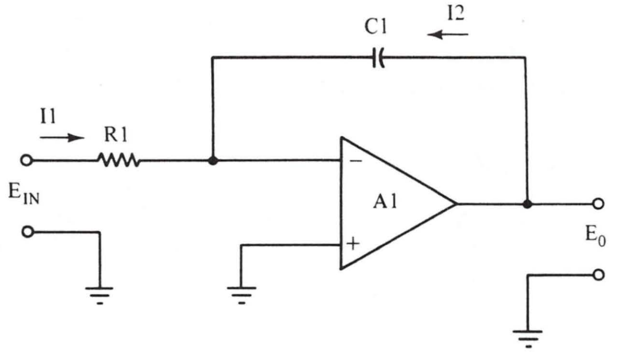


Figure 12-25. Integrator circuit.

Equation (12-41), then, is the transfer equation for the operational amplifier integrator circuit.

Example 12-4

A constant potential of 2 volts is applied to the input of the integrator in Fig. 12-19 for 3 seconds. Find the output potential if $R1 = 1$ megohm and $C1 = 0.5\ \mu\text{F}$.

SOLUTION

$$E_0 = \frac{-1}{R1C1} \int_0^t E_{\text{in}}\, dt \tag{12-42}$$

$$E_0 = \frac{-E_{\text{in}}}{R1C1} \int_0^3 dt$$

$$E_0 = \frac{(-2\text{ V})(t)}{(10^6\ \Omega)(5 \times 10^{27}\text{ F})}\Bigg|_0^3$$

$$E_0 = \frac{(-2\text{ V})(3\text{ s})}{(5 \times 10^{22}\text{ s})} - 0 = \mathbf{-12\ volts}$$

Note that the *gain* of the integrator is given by the term $1/R1C1$. If small values of $R1$ and $C1$ are used, then the gain can be very large. For example, if $R1$

= 100 kΩ, and $C1 = 0.001\ \mu F$, then the gain is 10,000. A very small input voltage in that case will saturate the output very quickly. In general, the *time constant* $R1C1$ should be longer than the period of the input waveform.

12-16 DIFFERENTIATORS

An operational amplifier *differentiator* is formed by reversing the roles of $R1$ and $C1$ in the integrator, as shown in Fig. 12-26. We know that

$$I2 = -I1 \tag{12-43}$$

$$I1 = \frac{C1\, dE_{in}}{dt} \tag{12-44}$$

and

$$I2 = E_0/R1 \tag{12-45}$$

Substituting Eqs. (12-44) and (12-45) into Eq. (12-43) results in

$$\frac{E_0}{R1} = \frac{-C1\, dE_{in}}{dt} \tag{12-46}$$

Solving Eq. (12-46) for E_o gives us the transfer equation for an operational amplifier differentiator circuit:

$$E_0 = -R1C1\,\frac{dE_{in}}{dt} \tag{12-47}$$

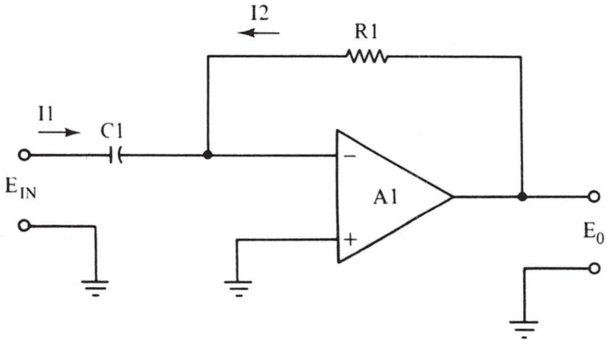


Figure 12-26. Differentiator circuit.

Example 12-5

A 12-volt/second ramp function voltage is applied to the input of an operational amplifier differentiator, in which $R1 = 1$ megohm and $C1 = 0.2\ \mu\text{F}$. What is the output voltage?

SOLUTION

$$E_0 = -R1C1(dE_{\text{in}}/dt) \tag{12-48}$$

$$E_0 = -(10^6\ \Omega)(2 \times 10^{-7}\ \text{F})(12\ \text{V/s})$$

$$E_0 = -(2 \times 10^{-1}\ \text{s})(12\ \text{V/s}) = \mathbf{-2.4\ V}$$

The differentiator time constant $R1C1$ should be set very short relative to the period of the waveform being differentiated, or in the case of square waves, triangle waves, and certain other signals, the time constant should be short compared with the *rise time* of the leading edge.

12-17 LOGARITHMIC AND ANTILOG AMPLIFIERS

Figure 12-27(a) shows an elementary *logarithmic amplifier* circuit using a bipolar transistor in the feedback loop. We know that the collector current bears a logarithmic relationship to the base-emitter potential, V_{be}:

$$V_{\text{be}} = \frac{KT}{q} \ln \left[\frac{I_c}{I_s}\right] \tag{12-49}$$

where V_{be} = the base-emitter potential in volts (V)
K = Boltzmann's constant 1.38×10^{-23} joules/degree Kelvin (j/°K)
T = the temperature in degrees Kelvin (°K)
q = the electronic charge (1.6×10^{-19} coulombs)
I_c = the collector current in amperes (A)
I_s = the reverse saturation current for the transistor in amperes (A)

At 27°C (300°K), i.e., room temperature, the term KT/q evaluates to approximately 26 mV (i.e., 0.026 V), so Eq. (12-49) becomes

$$V_{\text{be}} = 26\ \text{mV} \ln (I_c/I_s) \tag{12-50}$$

But $V_{\text{be}} = E_0$, and $I_c = E_{\text{in}}/R1$, so we may safely say that

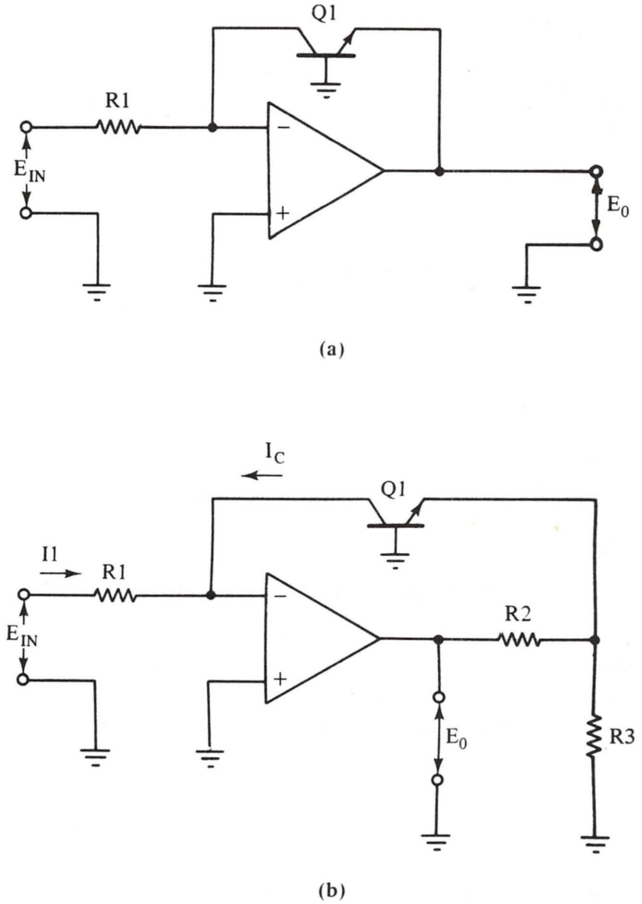


Figure 12-27. (a) Logarithmic amplifier. (b) Improved logarithmic amplifier.

$$E_0 = 26\text{ mV} \ln \left[\frac{E_{\text{in}}}{I_s R1} \right] \tag{12-51}$$

But I_s is a constant if the temperature is also constant, and $R1$ is constant under all conditions, so, by the rule that the logarithm of a constant is also a constant, we may state that Eq. (12-51) is the transfer function of the natural logarithmic amplifier. For base-10 logarithms:

$$E_0 = 60 \text{ mV} \log_{10} \left[\frac{E_{\text{in}}}{I_s R1} \right] \tag{12-52}$$

The relationship of Eqs. (12-51) and (12-52) allows us to construct amplifiers with logarithmic properties. If $Q1$ is in the feedback loop of an operational amplifier, then the output voltage E_o will be proportional to the logarithm of input voltage E_{in}. If, on the other hand, the transistor is connected in series with the input of the operational amplifier (see Fig. 12-28), then the circuit becomes an antilog amplifier.

Both of these circuits exhibit a strong dependence on temperature, as evidenced by the T term in Eq. (12-49). In actual practice, then, some form of temperature correction must be used. Two forms of temperature correction are commonly used: *compensation* and *stabilization*.

The compensation method uses temperature-dependent resistors, i.e., *thermistors*, to regulate the gain of the circuit with changes in temperature. For example, it is common practice to make $R3$ in Fig. 12-27(b) a thermistor.

The stabilization method requires that the temperature of $Q1$, and preferably the op-amp also, be held constant. In the past, this has meant that the components must be kept inside an electrically heated oven, but today other techniques are used. One manufacturer builds a temperature-controlled hybrid logarithmic amplifier by nesting the op-amp and transistor on the same substrate as a class-A amplifier. Such an amplifier, under zero-signal conditions, dissipates very nearly constant heat. After the chip comes to equilibrium, the temperature will remain constant.

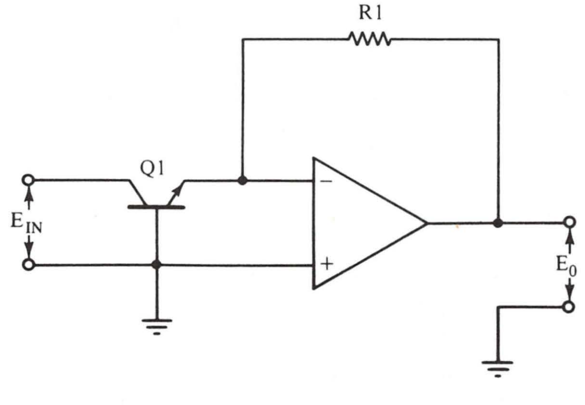


Figure 12-28. Antilog amplifier.

In the case of the antilog amplifier:

$$I_c = E_0/R1 \tag{12-53}$$

and

$$E_{in} = V_{be} \tag{12-54}$$

$$E_{in} = 26\ \text{mV}\ \ln\,(E_0/R1I_s) \tag{12-55}$$

12-18 CURRENT-TO-VOLTAGE CONVERTERS

Most analog recording devices, such as oscilloscopes or graphic recorders, are voltage-input devices. That is to say, they require a *voltage* for an input signal. When measuring or recording a current, however, they require some sort of *current-to-voltage* converter circuit.

Two examples of operational amplifier versions are shown in Fig. 12-29. In the first example, Fig. 12-29(a), a small value resistor *R* is placed in series with the current *I*, which produces a voltage equal to *IR*. This potential is seen by the operational amplifier as a valid input voltage. The output voltage is

$$E_0 = \frac{-IRR_f}{R_{in}}$$

provided that

$$R << R_{in}$$
$$R << R_f$$

The circuit shown in Fig. 12-23(b) is used for small currents. The output voltage in that circuit is given by

$$E_0 = -I_{in}R_f \tag{12-56}$$

The circuit shown in Fig. 12-29(b) is used for small currents.

12-19 SUMMARY

1. The transfer function of an operational amplifier circuit can be determined by the basic properties, Ohm's law, and Kirchoff's law.
2. Operational amplifiers can be used as amplifiers, integrators, differentiators, and so forth, to provide numerous functions, both linear and nonlinear.
3. Operational amplifiers can be configured to provide differential inputs.
4. The gain of an operational amplifier circuit can be set by the ratio of two resistors.

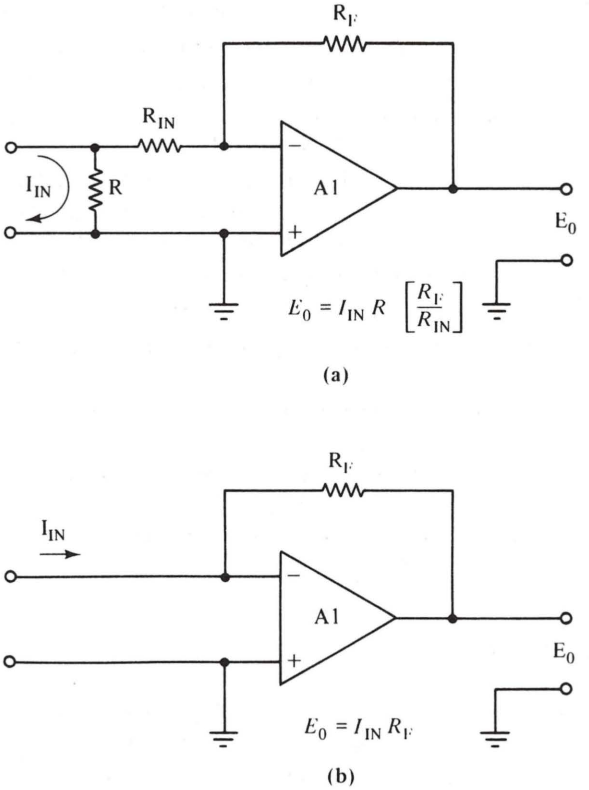


Figure 12-29. (a) Current-to-voltage converter. (b) Current-to-voltage converter for small currents.

12-20 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside any problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

12-21 QUESTIONS

1. Write the transfer function for an inverting follower.
2. Write the transfer function for a noninverting follower.
3. Write the transfer function for a differential amplifier.
4. Write the transfer function for an instrumentation amplifier.

5. Write the transfer function for an operational amplifier integrator.
6. Write the transfer function for an operational amplifier differentiator.
7. State four of the basic properties of an ideal operational amplifier.
8. Describe in your own words "virtual ground." State the property of ideal operational amplifiers on which this term is based.
9. An operational amplifier is designed to operate from two power supplies. V_{cc} is ____________ with respect to ground, while V_{ee} is ____________ with respect to ground.
10. Name three mechanisms by which an output offset voltage is created.
11. List two different methods for eliminating the output offset voltage.
12. The common mode voltage gain for a perfect operational amplifier is ____________.
13. The gain of an operational amplifier integrator is given by the expression: ____________________.
14. Write the transfer equations for (a) base-*e*, (b) base-10 logarithmic amplifiers.
15. What steps may be taken to reduce the effect of temperature on the logarithmic amplifier?

12-22 PROBLEMS

1. Calculate the voltage gain of an inverting follower if the feedback resistor is 560 kΩ, and the input resistor is 82 kΩ.
2. Calculate the voltage gain of an inverting follower if the feedback resistor is 100 kΩ and the input resistor is 10 kΩ.
3. Calculate the voltage gain of an inverting follower if the feedback resistor is 5.6 kΩ and the input resistor is 560 ohms.
4. Calculate the voltage gain of an inverting follower if the feedback resistor is 5 kΩ and the input resistor is 10 kΩ.
5. Calculate the voltage gain of a noninverting follower if the feedback resistor is 120 kΩ and the input resistor is 56 kΩ.
6. Calculate the voltage gain of a noninverting follower if the feedback resistor is 10 kΩ and the input resistor is 10 kΩ.
7. Calculate the voltage gain of a noninverting follower if the feedback resistor is 0 ohms and the input resistor is 100 kΩ.
8. Calculate the voltage gain of a noninverting follower if the feedback resistor is 2.2 kΩ and the input resistor is 4.7 kΩ.

In Problems 9 through 11 refer to Fig. 12-30.

9. In Fig. 12-30 point A is at a potential of +1 volt, point B is at +6 volts, and point D is at 0 volts. Find the output voltage if $R2 = 10\text{k}\Omega$ and $R1 = 9.1$ kΩ.

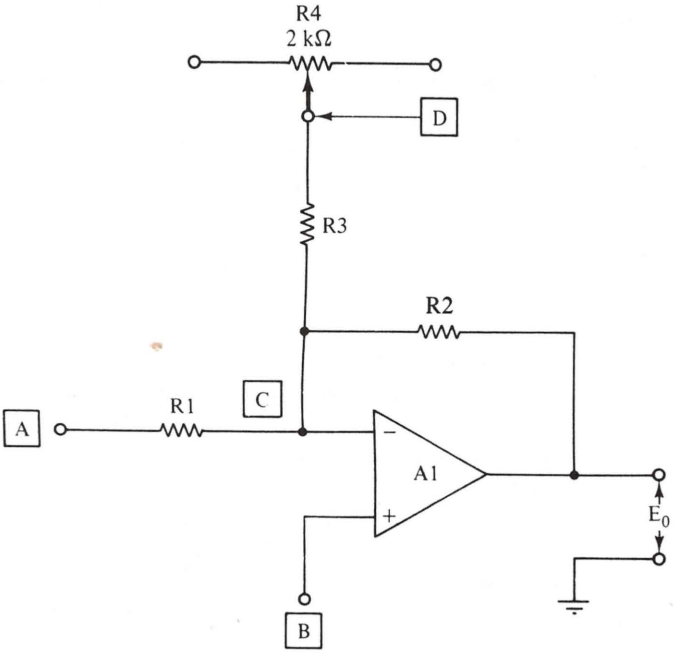


Figure 12-30.

10. In Fig. 12-30 point A is grounded, point B is at a dc potential of −3 volts, and point D is at a potential of +1.5 volts dc. Find the output potential if $R1 = R2 = 100\text{k}\Omega$ and $R3 = R1/2$.
11. In Fig. 12-30 points A and B are grounded, and point D is at a potential of −1 volt dc. Find $R2$ if $R3 = 2.2\ \text{k}\Omega$ and $E_o = +10$ volts.
12. Find the gain of an instrumentation amplifier such as Fig. 12-10 if $R1 =$ 500 ohms, $R2 = 16\ \text{k}\Omega$, $R4 = 3.9\ \text{k}\Omega$, and $R7 = 27\ \text{k}\Omega$.
13. An operational amplifier integrator uses a 0.69-μF capacitor and an 820-kΩ resistor. What will the output voltage be if a 100-mV dc level is applied to the input for 4 seconds?
14. What is the gain of an integrator that uses a 0.1-μF capacitor and a 47-kΩ resistor?
15. An operational amplifier differentiator uses a 1-megohm resistor and a 0.68-μF capacitor. What output voltage exists when an 8 V/sec ramp function is applied to the input?
16. Find V_{be} for a transistor operated at 40°C if the reverse saturation current at that temperature is 10^{-12} amperes, and the collector current is 1.2 mA.
17. A logarithmic amplifier is adjusted for the natural log operation mode. Find the output voltage if the reverse saturation current is 10^{-13} amperes, the input resistor is 100 kΩ, and $E_{in} = 100$ mV.

18. What is the *slope factor* of a natural logarithmic amplifier at 37°C?

19. A current-to-voltage converter consists of an operational amplifier and a 100-kΩ resistor in the negative feedback loop. Find E_0 if the input current is 20μA.

20. Derive the transfer function for an antilog amplifier. Do not refer back to the text.

12-23 REFERENCES

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chapter 13

TRANSDUCERS

13-1 OBJECTIVES

1. To learn the principles behind the operation of common types of transducer.
2. To learn how to specify and apply transducers.
3. To learn the limitations of certain transducers.
4. To learn about certain application problems and how they are solved.

13-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Describe the *differences* between *bonded* and *unbonded* piezoresistive strain gages.
2. List three different types of temperature transducer.
3. List two types of gas flow transducer.
4. Define "transducer" in your own words.
5. How may a force be measured if only displacement transducers are available?

13-3 TRANSDUCERS AND TRANSDUCTION

Not all of the physical variables that must be measured lend themselves to direct input into electronic instruments and circuits. Unfortunately, electronic circuits operate only with inputs that are currents and voltages. So, when one is measuring nonelectrical physical quantities it becomes necessary to provide a device that converts physical parameters such as *force, displacement, temperature, etc.*, into proportional voltages or currents. The transducer is such a device.

> *Definition:* A *transducer* is a device or apparatus that converts nonelectrical physical parameters into electrical signals, i.e., currents or voltages, that are proportional to the value of the physical parameter being measured.

Transducers take many forms, and may be based on a wide variety of physical phenomena. Even when one is measuring the *same* parameter, different instruments may use different types of transducer.

This chapter will not be an exhaustive catalogue treatment covering all transducers—the manufacturer's data sheets may be used for that purpose—but we will discuss some of the more common *types* of transducer used in scientific, industrial, medical, and engineering applications.

13-4 THE WHEATSTONE BRIDGE

Many forms of transducer create a variation in an electrical resistance, inductance, or capacitance in response to some physical parameter. These transducers are often in the form of a Wheatstone bridge, or one of the related ac bridge circuits. In many cases where the transducer itself is not in the form of a bridge, it is used in a bridge circuit with other components forming the other arms of the bridge. At this point, then, it may be advisable for the reader to quickly review Chapter 3, "AC and DC Bridges."

13-5 STRAIN GAGES

All electrical conductors possess some amount of electrical resistance. A bar or wire made of such a conductor will have an electrical resistance that is given by

$$R = \rho(L/A) \tag{13-1}$$

where R = the resistance in ohms (Ω)
ρ = the *resistivity constant,* a property specific to the conductor material, given in units of ohm-centimeters (Ω-cm)
L = the length in centimeters (cm)
A = the cross-sectional area in square centimeters (cm^2)

Example 13-1

A constantan (i.e., 55 percent copper, 45 percent nickel) round wire is 10 cm long and has a radius of 0.01 mm. Find the electrical resistance in ohms. (*Hint:* The resistivity of constantan is 44.2×10^{-6} Ω-cm).

SOLUTION

$$R = \rho(L/A) \qquad (13\text{-}1)$$

$$R = \frac{(44.2 \times 10^{-6}\ \text{ohm-cm})(10\ \text{cm})}{\pi \left[0.01\ \text{mm} \times \dfrac{1\ \text{cm}}{10\ \text{mm}}\right]^2}$$

$$R = \frac{(4.42 \times 10^{-4}\ \text{ohm-cm}^2)}{\pi(0.001\ \text{cm})^2}$$

$$R = \frac{(4.42 \times 10^{-4}\ \text{ohm-cm}^2)}{\pi 10^{-6}\ \text{cm}^2} = \mathbf{141\ ohms}$$

Note that the resistivity factor (ρ) in Eq. (13-1) is a constant, so if length L or area A can be made to vary under the influence of an outside parameter, then the electrical resistance of the wire will change. This phenomenon is called *piezoresistivity,* and is an example of a transducible property of a material.

> *Definition: Piezoresistivity* is the change in the electrical resistance of a conductor due to changes in length and cross-sectional area. In piezoresistive materials mechanical deformation of the material produces changes in electrical resistance.

Figure 13-1 shows how an electrical conductor can use the piezoresistivity property to measure *strain,* i.e., *forces* applied to it in *compression* or *tension.* In Fig. 13-1(a) we have a conductor at rest, in which no forces are acting. The length is given as L_0 and the cross-sectional area as A_0. The resistance of this conductor, from Eq. (13-1), is

$$R_0 = \rho(L_0/A_0) \qquad (13\text{-}2)$$

where ρ = the resistivity as defined previously
R_0 = the resistance in ohms (Ω) when no forces are applied
L_0 = the resting, i.e., no force, length in cm
A_0 = the resting cross-sectional area in cm^2

But in Fig. 13-1(b) we see the situation where a compression force of magnitude F is applied along the axis in the inward direction. The conductor will deform, causing the length L_1 to decrease to $(L_0 - \Delta L)$, and the cross-sectional area to increase to $(A_0 + \Delta A)$. The electrical resistance decreases to $(R_0 - \Delta R)$:

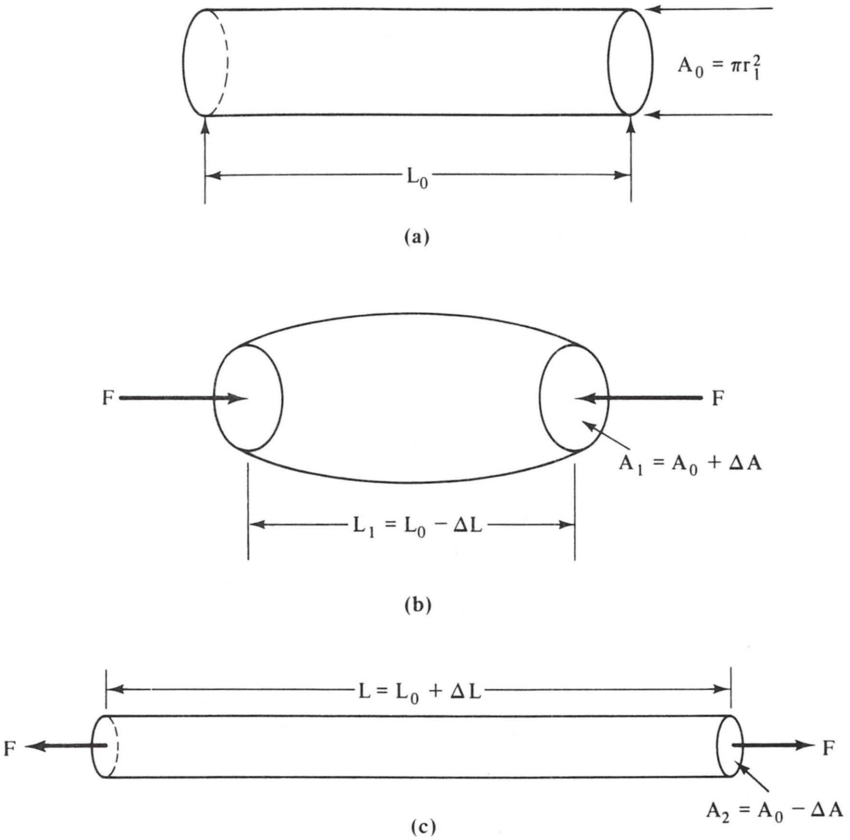


Figure 13-1. (a) Unstrained metal bar. (b) Metal bar in compression. (c) Metal bar in tension.

$$R_1 = (R_0 - \Delta R) \propto \frac{(L_0 - \Delta L)}{(A_0 + \Delta A)} \tag{13-3}$$

Similarly, when a tension force of the same magnitude (i.e., F) is applied—i.e., a force that is directed outward along the axis—the length increases to $(L_0 + \Delta L)$, and the cross-sectional area decreases to $(A_0 - \Delta A)$. The resistance will increase to

$$R_2 = (R_0 - \Delta R) \propto \frac{(L_0 + \Delta L)}{(A_0 + \Delta A)} \tag{13-4}$$

The *sensitivity* of the strain gage is expressed in terms of unit change of electrical resistance for a unit change in length, and is given in the form of a *gage factor S:*

$$S = \frac{(\Delta R/R)}{(\Delta L/L)} \tag{13-5}$$

where S = the gage factor (dimensionless)
R = the unstrained resistance of the conductor
ΔR = the change in resistance due to strain
L = the unstrained length of the conductor
ΔL = the change in length due to strain

Example 13-2

Find the gage factor of a 128-ohm conductor that is 24 mm long, if the resistance changes 13.3 ohms and the length changes 1.6 mm under a tension force.

SOLUTION

$$S = (\Delta R/R)/(\Delta L/L) \tag{13-5}$$

$$S = \frac{(13.3\ \Omega/128\ \Omega)}{(1.6\ \text{mm}/24\ \text{mm})}$$

$$S = \frac{1.04 \times 10^{-1}}{6.67 \times 10^{-2}} = \mathbf{1.56}$$

We may also express the gage factor in terms of the length and diameter of the conductor. Recall that the diameter is related to the cross-sectional area (i.e., $A = \pi d^2/4 = \pi r^2$), so the relationship between the gage factor S and these other factors is given by

$$S = 1 + 2\ \frac{(\Delta d/d)}{(\Delta L/L)} \tag{13-6}$$

Example 13-3

Calculate the gage factor S if a 1.5-mm-diameter conductor that is 24 mm long changes length by 1 mm and diameter by 0.02 mm under a compression force.

SOLUTION

$$S = 1 + 2\,\frac{(\Delta d/d)}{(\Delta d/d)} \tag{13-6}$$

$$S = 1 + \frac{2[(0.02\text{ mm})/(1.5\text{ mm})]}{(1\text{ mm})/(24\text{ mm})}$$

$$S = 1 + \frac{(2)(1.3 \times 10^{-2})}{(4.2 \times 10^{-2})}$$

$$S = 1 + (2)(0.31) = \mathbf{1.62}$$

Note that the expression $(\Delta L/L)$ is sometimes denoted by the Greek letter ϵ, so Eqs. (13-5) and (13-6) become

$$S = 1 + \frac{2(\Delta d/d)}{\epsilon}$$

$$S = \frac{(\Delta R/R)}{\epsilon}$$

Gage factors for various metals vary considerably. Constantan, for example, has a gage factor of approximately 2, while certain other common alloys have gage factors between 1 and 2. At least one alloy (92 percent platinum, 8 percent tungsten) has a gage factor of 4. Semiconductor materials such as germanium and silicon can be doped with impurities to provide custom gage factors between 50 and 250. The problem with semiconductor strain gages, however, is that they exhibit a marked sensitivity to temperature changes. Where semiconductor strain gages are used, either a thermally controlled environment or temperature compensating circuitry must be provided.

13-6 BONDED AND UNBONDED STRAIN GAGES

Strain gages can be classified as *unbonded* or *bonded*. These categories refer to the method of construction used. Figure 13-2 shows both methods of construction.

The unbonded type of strain gage is shown in Fig. 13-2(a), and consists of a wire resistance element stretched taut between two flexible supports. These

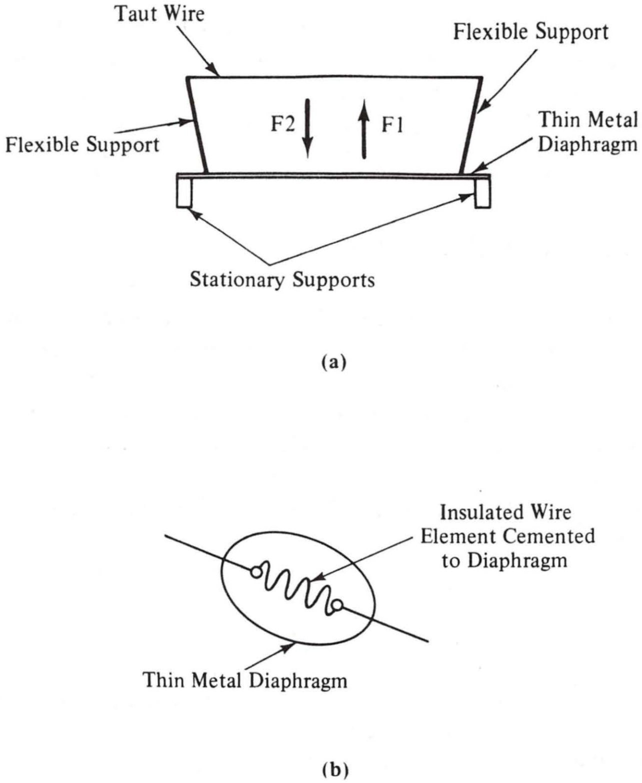


Figure 13-2. (a) Unbonded strain gage. (b) Bonded strain gage.

supports are configured in such a way as to place tension or compression forces on the taut wire when external forces are applied. In the particular example shown, the supports are mounted on a thin metal diaphragm that flexes when a force is applied. Force $F1$ will cause the flexible supports to spread apart, placing a tension force on the wire and increasing its resistance. Alternatively, when force $F2$ is applied, the ends of the flexible supports tend to move closer together, effectively placing a compression force on the wire element, thereby reducing its resistance. In actuality, the wire's resting condition is *tautness,* which implies a tension force. So $F1$ increases the tension force from normal, and $F2$ decreases the normal tension.

The bonded strain gage is shown in Fig. 13-2(b). In this type of device a wire or semiconductor element is cemented to a thin metal diaphragm. When the diaphragm is flexed, the element deforms to produce a resistance change.

The linearity of both types can be quite good, provided that the elastic limits of the diaphragm and the element are not exceeded. It is also necessary to insure that the ΔL term is only a very small percentage of L.

In the past it has been "standard wisdom"—i.e., the opinions of those who make purchase decisions—that bonded strain gages are more rugged, but less linear, than unbonded models. Although this may have been true at one time, recent experience has shown that modern manufacturing techniques produce linear, reliable instruments of both types.

13-7 STRAIN GAGE CIRCUITRY

Before a strain gage can be useful, it must be connected into a circuit that will convert its resistance changes to a current or voltage output. Most applications are voltage output circuits.

Figure 13-3(a) shows the *half-bridge* (so called because it is actually half of a Wheatstone bridge circuit) or *voltage divider* circuit. The strain gage element of resistance R is placed in series with a fixed resistance $R1$ across a stable and well-regulated voltage source E. The output voltage E_0 is found from the voltage divider equation

$$E_0 = \frac{ER}{R + R1} \tag{13-7}$$

Equation (13-7) describes the output voltage E_0 when the transducer is at rest, i.e., nothing is stimulating the strain gage element. When the element is stimulated, however, its resistance changes a small amount ΔR. To simplify our discussion we will adopt the standard convention used in many texts of letting $h = \Delta R$.

$$E_0 = \frac{E(R + h)}{(R \pm h) + R1} \tag{13-8}$$

Another half-bridge is shown in Fig. 13-3(b), but in this case the strain gage is in series with a *constant current source* (CCS), which will maintain current I at a constant level regardless of changes in strain gage resistance. The normal output voltage E_0 is

$$E_0 = IR \tag{13-9}$$

for nonstimulated conditions, and

$$E_0 = I(R \pm h) \tag{13-10}$$

under stimulated conditions.

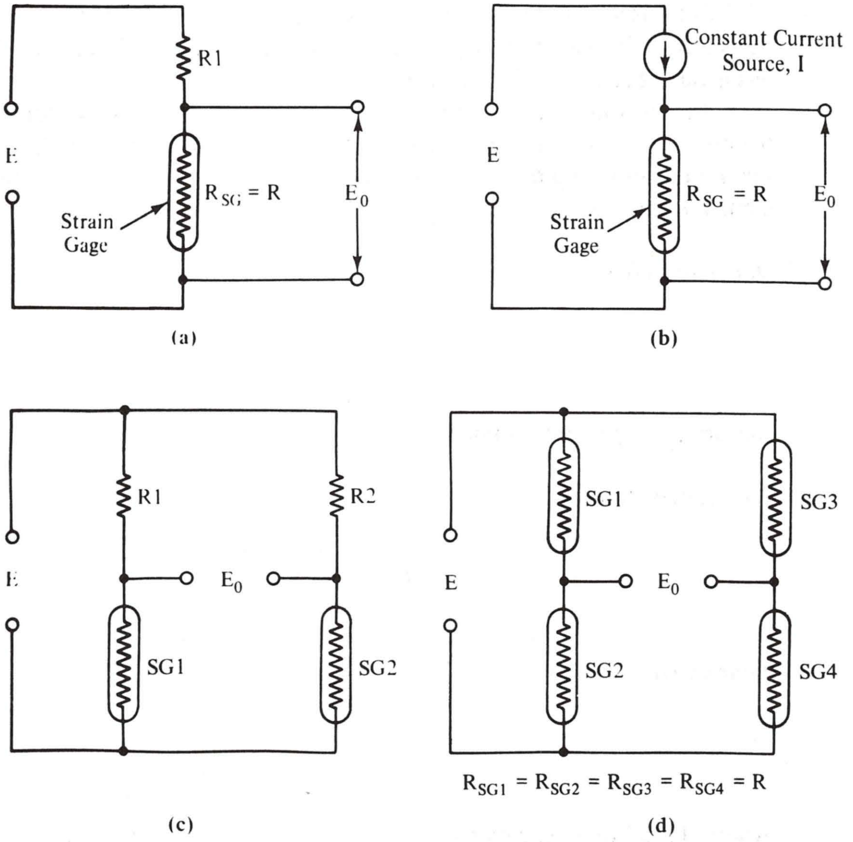


Figure 13-3. (a) Constant voltage strain gage circuit (half-bridge type). (b) Constant current strain gage circuit (half-bridge type). (c) Two strain -gage elements in Wheatstone bridge. (d) Four-active-element strain gage Wheatstone bridge.

The half-bridge circuits suffer from one major defect: Output voltage E_0 will always be present regardless of the stimulus. Ideally, in any transducer system, we want E_0 to be zero when the stimulus is also zero, and take a value proportional to the stimulus when the stimulus value is nonzero. A Wheatstone bridge circuit in which one or more strain gage elements form the bridge arms has this property.

Figure 13-3(c) shows a circuit in which strain gage elements $SG1$ and $SG2$ form two bridge arms and fixed resistors $R1$ and $R2$ form the other two arms. It is usually the case that $SG1$ and $SG2$ will be configured so that their actions oppose each other; that is, under stimulus, $SG1$ will have a resistance $R + h$ and $SG2$ will have a resistance $R - h$, or vice versa.

One of the most linear forms of transducer bridge is the circuit of Fig.

13-3(d) in which all four bridge arms contain strain gage elements. In most such transducers all four strain gage elements have the same resistance, i.e., R, which has a value between 100 ohms and 1000 ohms in most cases.

Recall from Chapter 3 that the output voltage from a Wheatstone bridge is the difference between the voltages across the two half-bridge dividers. The following equations hold true for bridges in which one, two, or four equal active elements are used.

One active element:

$$E_0 = \frac{E}{4}\left[\frac{h}{R}\right] \tag{13-11}$$

(accurate ±5 percent, provided that $h \leqslant 0.1$)

Two active elements:

$$E_0 = \frac{E}{2}\left[\frac{h}{R}\right] \tag{13-12}$$

Four active elements:

$$E_0 = \frac{Eh}{R} \tag{13-13}$$

where, for all three equations, E_0 = the output potential in volts (V)
E = the excitation potential in volts (V)
R = the resistance of all bridge arms
h = the quantity ΔR, the change in resistance of a bridge arm under stimulus

(These equations apply only for the case where all the bridge arms have equal resistances under zero stimulus conditions.)

Example 13-4

A transducer that measures force has a nominal resting resistance of 300 ohms, and is excited by +7.5 volts dc. When a 980-dyne force is applied, all four equal-resistance bridge elements change resistance by 5.2 ohms. Find the output voltage E_0.

SOLUTION

$$E_0 = E(h/R) \qquad (13\text{-}13)$$
$$E_0 = (7.5\ \text{V})(5.2\ \Omega/300\ \Omega)$$
$$E_0 = (7.5\ \text{V})(5.2)/(300) = \mathbf{0.13\ V}$$

13-8 TRANSDUCER SENSITIVITY (ψ)

When designing electronic instrumentation systems involving strain gage transducers, it is convenient to use the *sensitivity factor* (denoted by the Greek letter "psi"—ψ), which relates the output voltage in terms of the excitation voltage and the applied stimulus. In most cases, we see a specification giving the number of microvolts or millivolts output per volt of excitation potential per unit of applied stimulus (i.e., $\mu\text{V}/\text{V}/Q_0$ or $\text{mV}/\text{V}/Q_0$).

$$\psi = E_0'/V/Q_0 \qquad (13\text{-}14\text{a})$$

and

$$\psi = \frac{E_0'}{V \times Q_0} \qquad (13\text{-}14\text{b})$$

where E_0' = the output potential
V = one unit of potential, i.e., 1 volt
Q_0 = one unit of stimulus

The sensitivity is often given as a specification by the transducer manufacturer. From it we can predict output voltage for any level of stimulus and excitation potential. The output voltage, then, is found from

$$E_0 = \psi E Q \qquad (13\text{-}15)$$

where E_0 = the output potential in volts (V)
ψ = the sensitivity in $\mu\text{V}/\text{V}/Q_0$
E = the excitation potential in volts (V)
Q = the stimulus parameter

Example 13-5

A well-known medical arterial blood pressure transducer uses a four-element piezoresistive Wheatstone bridge with a sensitivity of 5 microvolts per volt of excitation per Torr of pressure, i.e., 5 $\mu\text{V}/\text{V}/\text{T}$ (*Note:* 1 Torr = 1 mm Hg). Find the output voltage if the bridge is excited by 5 volts dc and 120 Torr of pressure is applied.

SOLUTION

$$E_0 = \psi EQ \qquad (13\text{-}15)$$

$$E_0 = \frac{5\ \mu\text{V}}{\text{V} - \text{T}} \times (5\ \text{V}) \times (120\ \text{T})$$

$$E_0 = (5 \times 5 \times 120)\ \mu\text{V} = \mathbf{3000\ \mu V}$$

13-9 BALANCING AND CALIBRATING THE BRIDGE

Few, if any, Wheatstone bridge strain gages meet the ideal condition in which all four arms have exactly equal resistances. In fact, the bridge resistance specified by the manufacturer is a *nominal* value only. There will inevitably be an *offset voltage,* i.e., $E_0 \neq 0$ when $Q = 0$. Figure 13-4 shows a circuit that will balance the bridge when the stimulus is zero. Potentiometer $R1$, usually a type with ten or more turns of operation, is used to inject a balancing current I into the bridge circuit at one of the nodes. $R1$ is adjusted, with the stimulus at zero, for zero output voltage.

The best calibration method is to apply a precisely known value of stimulus to the transducer, and adjust the amplifier following the transducer for the output

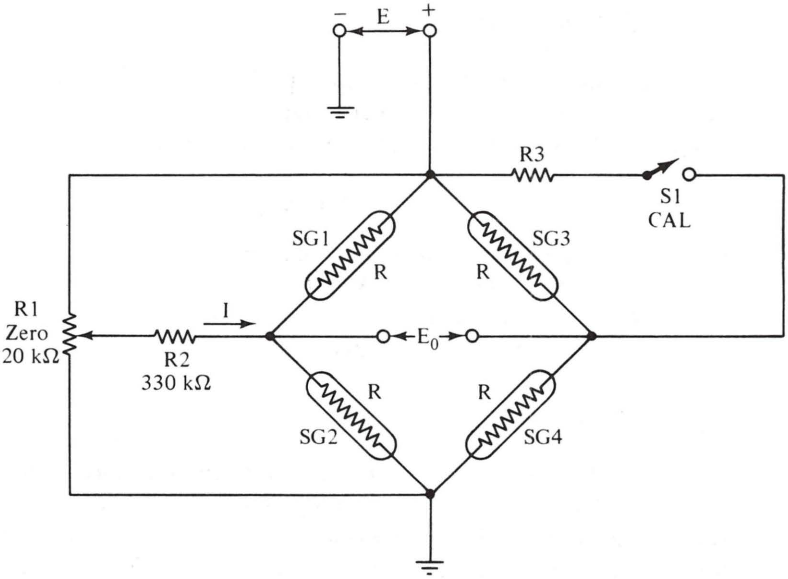


Figure 13-4. Circuit for using Wheatstone bridge transducer.

proper for that level of stimulus. But that may prove unreasonably difficult in some cases, so an *artificial* calibrator is needed to simulate the stimulus. This function is provided by $R3$ and $S1$ in Fig. 13-4. When $S1$ is open, the transducer is able to operate normally, but when $S1$ is closed it *unbalances* the bridge and produces an output voltage E_0 that simulates some standard value of the stimulus. The value of $R3$ is given by

$$R3 = \left[\frac{R}{4Q\psi} - \frac{R}{2}\right] \qquad (13\text{-}16)$$

where $R3$ = the resistance of R3 in ohms
R = the nominal resistance of the bridge arms in ohms
Q = the calibrated stimulus parameter
ψ = the sensitivity factor in V/V/Q (*Note* the difference in the units of Ψ: V instead of μV)

Example 13-6

An arterial blood pressure transducer has a sensitivity of 10 μV/V/Torr, and a nominal bridge arm resistance of 200 ohms. Find a value for $R3$ in Fig. 13-4 if we want the calibration to simulate an arterial pressure of 200 mm Hg (i.e., 200 Torr).

SOLUTION

$$R3 = \left[\frac{R}{4Q\psi} - \frac{R}{2}\right] \qquad (13\text{-}16)$$

$$R3 = \frac{200\ \Omega}{4 \times \left(\dfrac{10^{-5}\ \text{V}}{\text{V}-\text{T}}\right) \times 200\ \text{T}} - \frac{200\ \Omega}{2}$$

$$R3 = \frac{200\ \Omega}{(4)(10^{-5})(200)} - 100\ \Omega = \mathbf{24{,}900\ \Omega}$$

13-10 TEMPERATURE TRANSDUCERS

A large number of physical phenomena are temperature dependent, so we find quite a variety of electrical temperature transducers on the market. In this discussion, however, we will discuss only three basic types: *thermistor, thermocouple,* and *semiconductor pn junctions.*

13-11 THERMISTORS

Metals and most other conductors are temperature sensitive, and will change electrical resistance with changes in temperature, namely:

$$R_t = R_0[1 + \alpha(T - T_0)] \tag{13-17}$$

where R_t = the resistance in ohms at temperature T
R_0 = the resistance in ohms at temperature T_0 (often a standard reference temperature)
T = the temperature of the conductor
T_0 = a previous temperature of the conductor at which R_0 was determined.
α = the *temperature coefficient* of the material, a property of the conductor, in °C^{-1}

The temperature coefficients of most metals is positive, as are the coefficients for most semiconductors; e.g., gold has a value of +0.004/°C. Ceramic semiconductors used to make *thermistors,* i.e., thermal resistors, can have either negative or positive temperature coefficients depending upon their composition.

The resistance of a thermistor is given by

$$R_t = R_0 e^{\beta[(1/T) - 1/T_0)]}$$

where R_t = the resistance of the thermistor at temperature T
R_0 = the resistance of the thermistor at a reference temperature (usually the ice point, 0°C, or room temperature, 25°C)
e = the base of the natural logarithms
T = the thermistor temperature in degrees Kelvin (°K)
T_0 = the reference temperature in degrees Kelvin (°K)
β = a property of the material used to make the thermistor

(*Note:* β will usually have a value between 1500°K and 7000°K.)

Example 13-7

Calculate the resistance of a thermistor at 100°C if the resistance at 0°C was 18 kΩ. The material of the thermistor has a value of 2200°K. (*Note:* 0°C = 273°K, so 100°C = 373°K.)

SOLUTION

$$R_t = R_0 \exp \beta \left(\frac{1}{T} - \frac{1}{T_0}\right) \tag{13-18}$$

$$R_t = (1.8 \times 10^4\,\Omega) \exp \left[(2200^\circ\mathrm{K}) \left(\frac{1}{373^\circ\mathrm{K}} - \frac{1}{273^\circ\mathrm{K}}\right)\right]$$

$$R_t = 1.8 \times 10^4 \;\Omega\; e^{-2.15} = \mathbf{2089} \;\Omega$$

Equation (13-18) demonstrates that the response of a thermistor is exponential, as shown in Fig. 13-5. Note that both curves are nearly linear over a portion of their ranges, but become decidedly nonlinear in the remainder of the region. If a wide measurement range is needed, then a linearization network will be required.

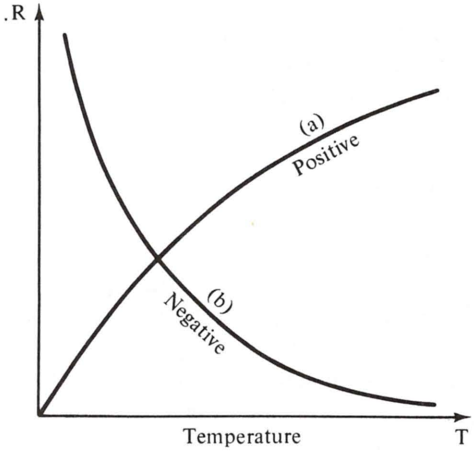


Figure 13-5. Thermistor temperature-vs-resistance curves.

Thermistor transducers will be used in any of the circuits in Fig. 13-3. They will also be found using many packaging arrangements. Figure 13-6 shows a bead thermistor used in medical instruments to continuously monitor a patient's rectal temperature.

The equations governing thermistors usually apply if there is little *self-heating* of the thermistor, although there are applications where self-heating is used. But in straight temperature measurements it is to be avoided. To minimize self-heating it is necessary to control the power dissipation of the thermistor.

Also of concern in some applications is the *time constant* of the thermistor. The resistance does not jump immediately to the new value when the temperature

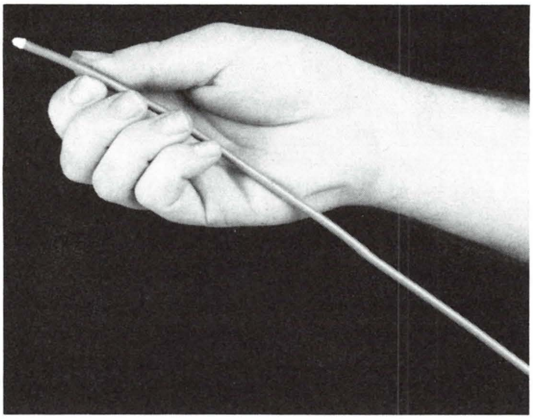


Figure 13-6. Thermistor in a medical rectal probe. (Courtesy of Electronics-for-Medicine.)

changes, but requires a small amount of time to stabilize at the new resistance value. This is expressed in terms of the time constant of the thermistor in a manner that is reminiscent of capacitors charging in RC circuits.

13-12 THERMOCOUPLES

When two dissimilar metals are joined together to form a "vee" [as in Fig. 13-7(a)], it is possible to generate an electrical potential merely by heating the junction. This phenomenon, first noted by Seebeck in 1823, is due to different *work functions* for the two metals. Such a junction is called a *thermocouple*. Seebeck EMF generated by the junction is proportional to the junction temperature, and is reasonably linear over wide temperature ranges.

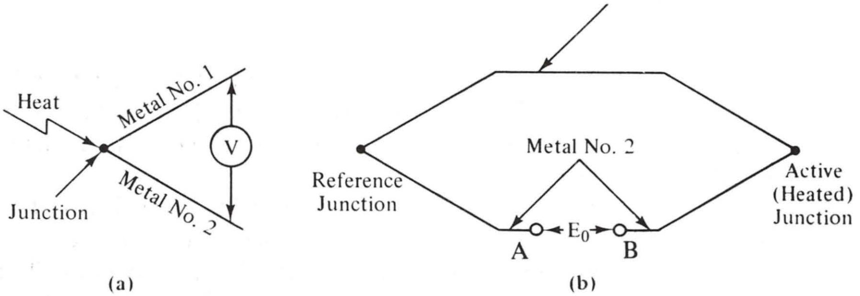


Figure 13-7. (a) Thermocouple junction. (b) Two-thermocouple temperature transducer.

A simple thermocouple is shown in Fig. 13-7(b), and uses two junctions. One junction is the measurement junction, and it is used as the thermometry probe. The other junction is a reference, and is kept at a reference temperature such as the ice point (0°C) or room temperature.

Interestingly enough, there is an inverse thermocouple phenomenon, called the *Peltier effect,* in which an electrical potential applied across A-B in Fig. 13-7(b) will cause one junction to absorb heat (i.e., get hot) and the other to lose heat (i.e., get cold). Semiconductor thermocouples have been used in small-scale environmental temperature chambers, and it is reported that one company researched the possibility of using Peltier devices to cool submarine equipment. Ordinary air conditioning equipment proves too noisy in submarines desirous of "silent running."

13-13 SEMICONDUCTOR TEMPERATURE TRANSDUCERS

Ordinary pn junction diodes exhibit a strong dependence upon temperature. This effect can be easily demonstrated by using an ohmmeter and an ordinary rectifier diode such as the 1N4000-series devices. Connect the ohmmeter so that it forward-biases the diode, and note the resistance at room temperature. Next hold a soldering iron or other heat source close to the diode's body, and watch the electrical resistance change. In a circuit such as Fig. 13-8 the current is held constant, so output voltage E_0 will change with temperature-caused changes in diode resistance.

Another solid-state temperature transducer is shown in Fig. 13-9. In this version, the temperature sensor device is a pair of diode-connected transistors. In any transistor the base-emitter voltage V_{be} is

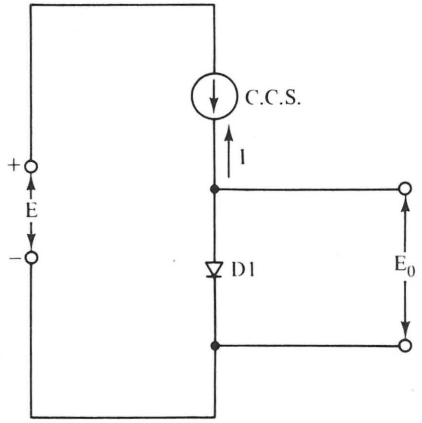


Figure 13-8. PN junction diode as a temperature transducer.

$$V_{be} = \frac{kT}{q} \ln\left[\frac{I_c}{I_s}\right] \tag{13-19}$$

where V_{be} = the base-emitter potential in volts (V)
k = Boltzmann's constant (1.38×10^{-23} J/°K)
T = the temperature in degrees Kelvin (°K)
Q = the electronic charge (1.6×10^{-19} coulomb)
ln denotes the natural logarithms
I_c = the collector current in amperes (A)
I_s = the reverse saturation current in amperes (A)

Note that the k and q terms in Eq. (13-19) are constants, and both currents can be made to be constant. The only variable, then, is *temperature*.

In the circuit of Fig. 13-9 we use two transistors connected to provide a differential output voltage ΔV_{be} that is the difference between $V_{be(Q1)}$ and $V_{be(Q2)}$. Combining the expressions for V_{be} for both transistors yields the expression

$$\Delta V_{be} = \frac{kT}{q} \ln\left[\frac{I1}{I2}\right] \tag{13-20}$$

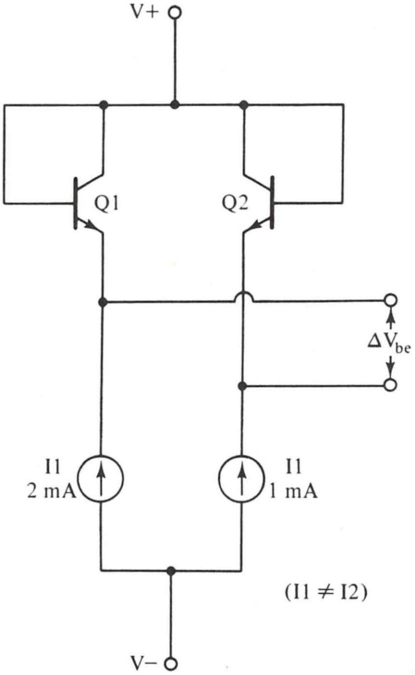


Figure 13-9. Two transistors connected as diodes from a temperature transducer if/1 =/2.

Note that, since $\ln 1 = 0$, currents $I1$ and $I2$ *must not be equal.* In general, designers set a ratio of 2:1, i.e., $I1 = 2$ mA and $I2 = 1$ mA. Since currents $I1$ and $I2$ are supplied from constant current sources, the ratio $I1/I2$ is a constant. Also, it is true that the logarithm of a constant is a constant. Therefore, all terms in Eq. (13-20) are constants, except temperature T. Equation (13-20), therefore, may be written in the form

$$\Delta V_{\text{be}} = KT \tag{13-21}$$

where $K = (k/q) \ln (I1/I2)$
$K = (1.38 \times 10^{-23})/(1.6 \times 10^{-19}) \ln (2/1)$
$K = 5.98 \times 10^{-5}\ \text{V/°K} = 59.8\ \mu\text{V/°}K$

We may now rewrite Eq. (13-21) in the form

$$\Delta V_{\text{be}} = 59.8\ \mu\text{V/°K}$$

Example 13-8

Calculate the output voltage from a circuit such as Fig. 13-9 if the temperature is 35°C (*Hint:* °K = °C + 273)

SOLUTION

$$\Delta V_{\text{be}} = KT \tag{13-21}$$

$$\Delta V_{\text{be}} = \frac{59.8\ \mu\text{V}}{°K} \times (35 + 273)°K$$

$$\Delta V_{\text{be}} = (59.8)(308)\ \mu\text{V} = 18{,}418\ \mu\text{V} = \mathbf{0.0184\ V}$$

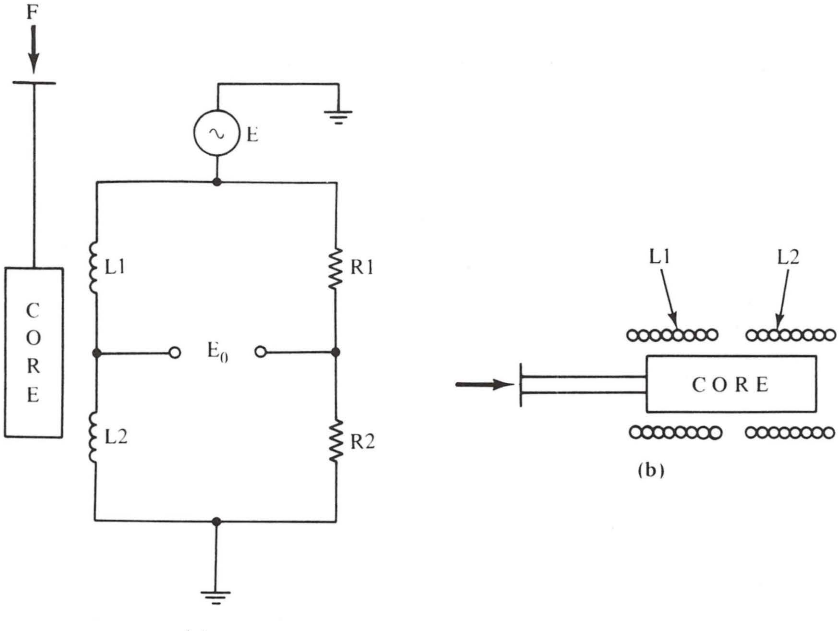
In most thermometers using the circuit of Fig. 13-9 an amplifier increases the output voltage to a level that is numerically the same as a unit of temperature, so that the temperature may be easily read from a digital voltmeter. The most common scale factor is 10 mV/°K, so for our transducer, the post-amplifier requires a gain of

$$A_{\text{v}} = \frac{10\ \text{mV/°K}}{59.8\ \mu\text{V} \times \dfrac{1\ \text{mV}}{10^3\ \mu\text{V}}} = 167$$

13-14 INDUCTIVE TRANSDUCERS

Inductance L and inductive reactance X_L are transducible properties, because they can be *varied* by certain mechanical methods.

Figure 13-10(a) shows an example of an inductive Wheatstone bridge. Resistors $R1$ and $R2$ form two fixed arms of the bridge, while coils $L1$ and $L2$ form variable arms. Since inductors are used, the excitation voltage must be ac. In most cases, the ac excitation source will have a frequency between 400 and 5000 Hz, and an rms amplitude of 5 to 10 volts.



(a)

(b)

Figure 13-10. (a) Inductive Wheatstone bridge transducer. (b) Mechanical form.

The inductors are constructed coaxially, as shown in Fig. 13-10(b), with a common core. It is a fundamental property of any inductor that a ferrous core increases its inductance. In the rest condition, i.e., zero-stimulus, the core will be positioned equally inside of both coils. If the stimulus moves the core in the direction shown in Fig. 13-10(b), the core tends to move out of $L1$ and further into $L2$. This action reduces the inductive reactance of $L1$ and increases that of $L2$, unbalancing the bridge.

13-15 LINEAR VARIABLE DIFFERENTIAL TRANSFORMERS (LVDT)

Another form of inductive transformer is the *linear variable differential transformer* (LVDT) shown in Fig. 13-11. The construction of the LVDT is similar to that of the inductive bridge, except that it also contains a primary winding.

One advantage of the LVDT over the bridge-type transducer is that it provides higher output voltages for small changes in core position. Several commer-

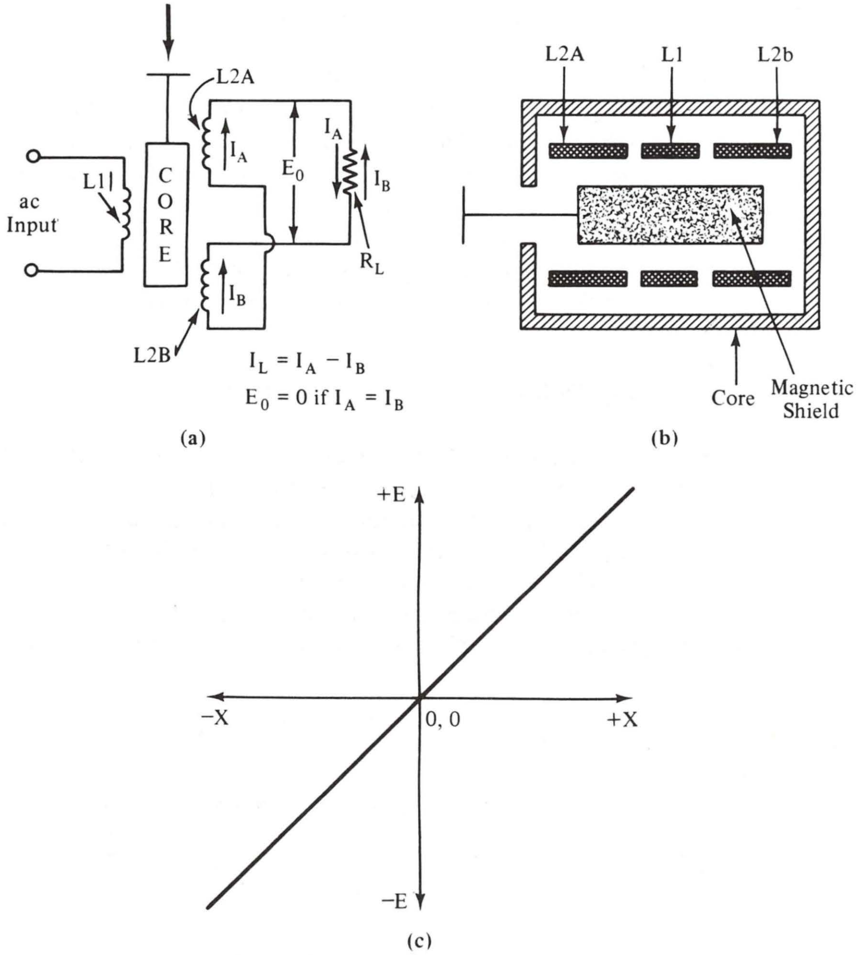


Figure 13-11. (a) LVDT. (b) LVDT construction. (c) Output transfer function.

cial models are available that produce 50 mV/mm to 300 mV/mm. In the latter case, this means that a 1-mm displacement of the core produces a voltage output of 300 mV.

In normal operation, the core is equally inside both secondary coils, *L2a* and *L2b,* and an ac carrier is applied to the primary winding. This carrier typically has a frequency between 40 Hz and 20 kHz, and an amplitude in the range 1 V rms and 10 V rms.

Under rest conditions the coupling between the primary and each secondary is equal. The currents flowing in each secondary, then, are equal to each other. Note in Fig. 13-11(a) that the secondary windings are connected in series-opposing, so if the secondary winding currents are equal, they will exactly cancel each other in the load. The ac voltage appearing across the load, therefore, is *zero* ($I_a = I_b$).

But when the core is moved so that it is more inside *L2b* and less inside *L2a*, the coupling between the primary and *L2b* is greater than the coupling between the primary and *L2a*. Since this fact makes the two secondary currents no longer equal, the cancellation is not complete. The current in the load I_L is no longer zero. The output voltage appearing across load resistor R_L is proportional to the core displacement, as shown in Fig. 13-11(c). The *magnitude* of the output voltage is proportional to the *amount* of core displacement, while the *phase* of the output voltage is determined by the *direction* of the displacement.

13-16 POSITION-DISPLACEMENT TRANSDUCERS

A position transducer will create an output signal that is proportional to the position of some object along a given axis. For very small position ranges we could use a strain gage (i.e., Fig. 13-12), but note that the range of such transducers is necessarily very small. Most strain gages either are nonlinear for large displacements, or are damaged by large displacements.

The LVDT can be used as a position transducer. Recall that the output polarity indicates the direction of movement from a zero-reference position, and the

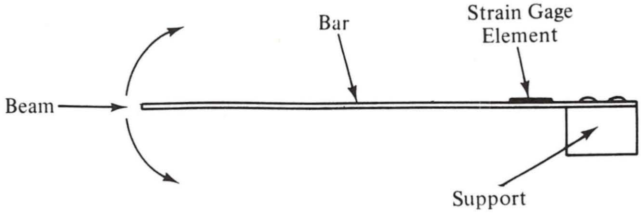


Figure 13-12. Beam transducer using a strain gage element.

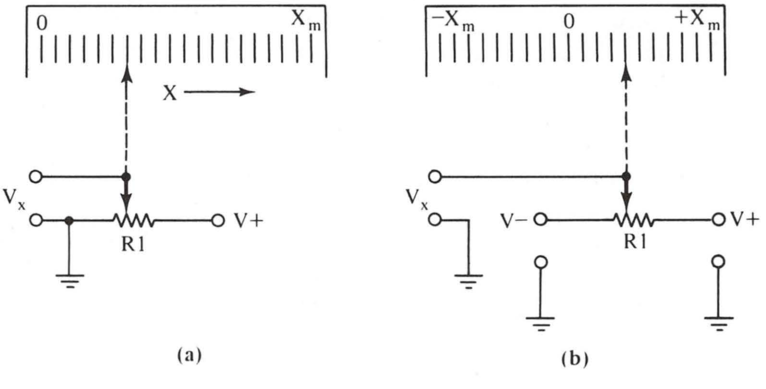
amplitude indicates the magnitude of the displacement. Although the LVDT will accommodate larger displacements than the strain gage, it is still limited in range.

The most common form of position transducer is the potentiometer. For applications that are not too critical, it is often the case that ordinary linear taper potentiometers are sufficient. Rotary models are used for curvilinear motion, and slide models for rectilinear motion.

We have already seen an example of the potentiometer as position transducers in our discussion (Chapter 10) on servomechanism strip-chart recorders.

In precision applications designers use either regular precision potentiometers or special potentiometers designed specifically as position transducers.

Figure 13-13 shows two possible circuits using potentiometers as position transducers. In Fig. 13-13(a) we see a single-quadrant circuit for use where the zero point, i.e., starting reference, is at one end of the scale. The pointer will always be at some point such that $0 \leqslant x \leqslant X_m$. The potentiometer is connected so that one end is grounded and the other is connected to a precision, regulated voltage source $V+$. The value of V_x represents X, and will be $0 \leqslant V_x \leqslant V+$, such that $V_x = 0$ when $X = 0$, and $V_x = V+$ when $X = X_m$.



(a) (b)

Figure 13-13. (a) Position transducer using a potentiometer (one quadrant). (b) Position transducer using a potentiometer (two quadrants).

A two-quadrant system is shown in Fig. 13-13(b), and is similar to the previous circuit except that instead of grounding one end of the potentiometer, it is connected to a precision, regulated *negative*-to-ground power source, $V-$. Figure 13-14 shows the output functions of these two transducers. Figure 13-14(a) represents the circuit of Fig. 13-13(a), while Fig. 13-14(b) represents the circuit of Fig. 13-13(b).

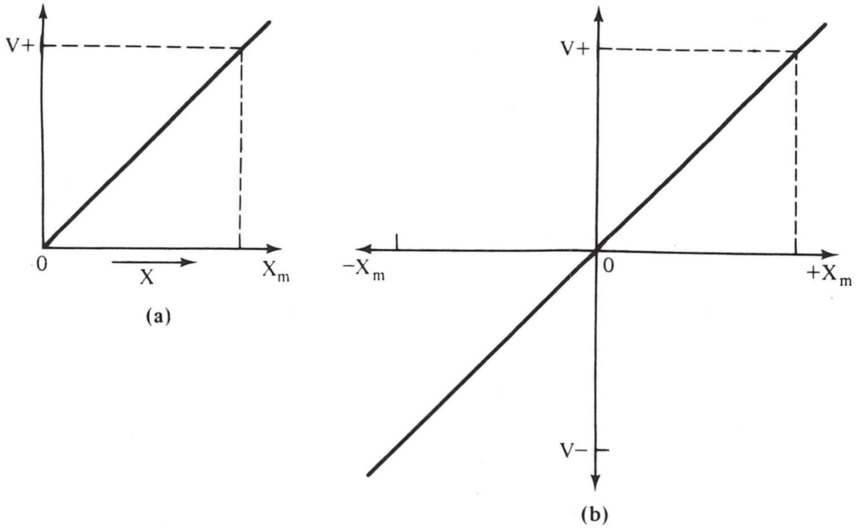


Figure 13-14. (a) *V+-vs-X* for Fig. 13-13(a). (b) *V-vs-X* for Fig. 13-13(b).

A four-quadrant transducer can be made by placing two circuits such as Fig. 13-13(b) at right angles to each other, and arranging linkage so that the output signal varies appropriately.

13-17 VELOCITY AND ACCELERATION TRANSDUCERS

Velocity can be defined as displacement per unit of time, and acceleration is the time rate of change of velocity. Since both velocity *(v)* and acceleration *(a)* can be related back to position *(s),* we often find position transducers used to *derive* velocity and acceleration signals. The relationships are

$$v = \frac{ds}{dt} \tag{13-22}$$

$$a = \frac{dv}{dt} \tag{13-23}$$

$$a = \frac{d^2s}{dt^2} \tag{13-24}$$

Velocity and acceleration are the first and second time derivatives of displacement (i.e., change of position), respectively. We may derive electrical signals proportional to *v* and *a* by using an operational amplifier differentiator circuit (see Fig. 13-15). The output of the transducer is a time-dependent function of position, i.e., displacement. This signal is differentiated by the stages following to produce the velocity and acceleration signals.

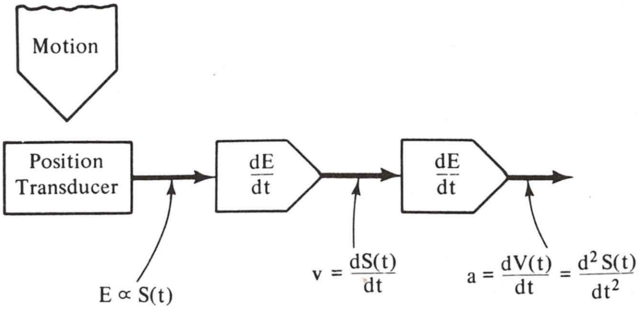


Figure 13-15. Example of derived signals using integration or differentiation.

13-18 TACHOMETERS

AC and dc generators are also used as velocity transducers. In their basic form they will transduce rotary motion—i.e., produce an angular velocity signal—but with appropriate mechanical linkage will also indicate rectilinear motion.

In the case of a dc generator, the output signal is a dc voltage with a magnitude that is proportional to the angular velocity of the armature shaft.

The ac generator, or *alternator*, maintains a relatively constant output voltage, but its ac *frequency* is proportional to the angular velocity of the armature shaft.

If a dc output is desired, instead of an ac signal, then a circuit similar to Fig. 13-16 is used. The ac output of the tachometer is fed to a trigger circuit (either a comparator or Schmitt trigger) so that squared-off pulses are created.

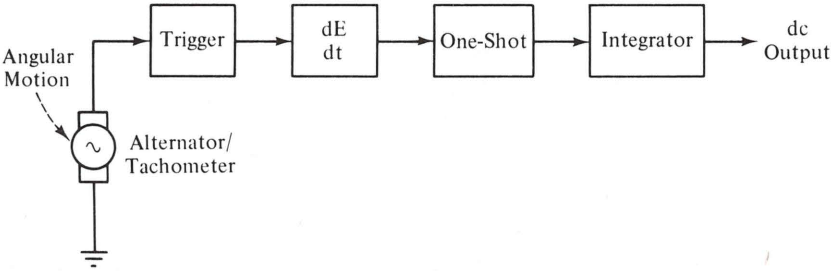


Figure 13-16. Alternator tachometer.

These pulses are then differentiated to produce spike-like pulses to trigger the monostable multivibrator (one-shot). The output of the one-shot is integrated to produce a dc level proportional to the tachometer frequency.

The reason for using the one-shot stage is to produce output pulses that have a *constant amplitude* and *duration*. Only the pulse repetition rate (i.e., number of pulses per unit of time) varies with the input frequency. This fact allows us to integrate the one-shot output to obtain our needed dc signal. If either duration or amplitude varied, then the integrator output would be meaningless. This technique, incidentally, is widespread in electronic instruments, so should be understood well.

13-19 FORCE AND PRESSURE TRANSDUCERS

Force transducers can be made by using strain gages, or either LVDT or potentiometer displacement transducers. In the case of the displacement transducer [Fig. 13-17(a)] it becomes a force transducer by causing a power spring to either

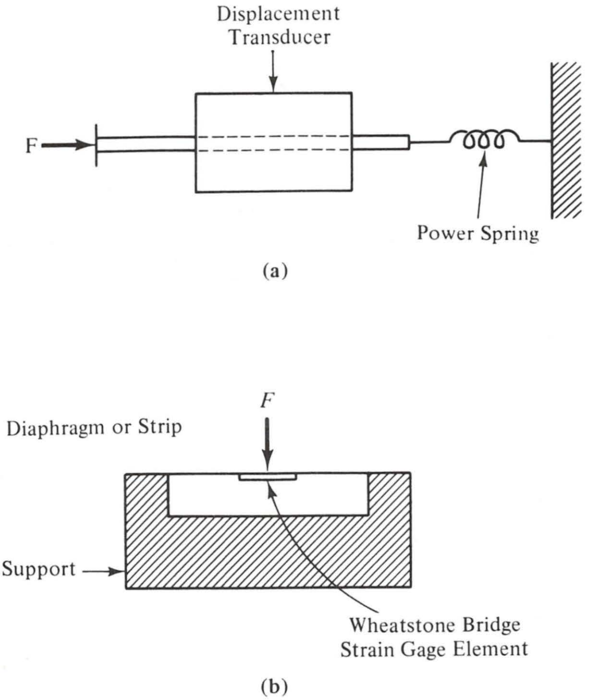


Figure 13-17. (a) Force from a displacement transducer. (b) Force/pressure transducer.

compress or stretch. Recall Hooke's law, which tells us that the force required to compress or stretch a spring is proportional to a *constant* and the *displacement* caused by the compression or tension force applied to the spring. So by using a displacement transducer and a calibrated spring, we are able to measure force.

Strain gages connected to flexible metal bars (refer back to Fig. 13-12) are also used to measure force, because it requires a certain amount of force to deflect the bar any given amount. There are several transducers on the market that use this technique, and they are advertised as "force-displacement" transducers. Such transducers form the basis of the digital bathroom scales now on the market.

Do not be surprised to see such transducers, especially the smaller types, calibrated in *grams*. We all know that the gram is a unit of *mass,* not force, so what this usage refers to is the gravitational force on one gram at the earth's surface, roughly 980 dynes. A 1-g weight suspended from the end of the bar in Fig. 13-12 will represent a force of 980 dynes.

A side view of a cantilever force transducer is shown in Fig. 13-17(b). In this device a flexible strip is supported by mounts at either end, and a piezoresistive strain gage is mounted to the under side of the strip. Flexing the strip unbalances the gage's Wheatstone bridge, producing an output voltage.

A related device uses a cup- or barrel-shaped support, and a circular diaphragm instead of the strip. Such a device will measure force or pressure, i.e., *force per unit of area.*

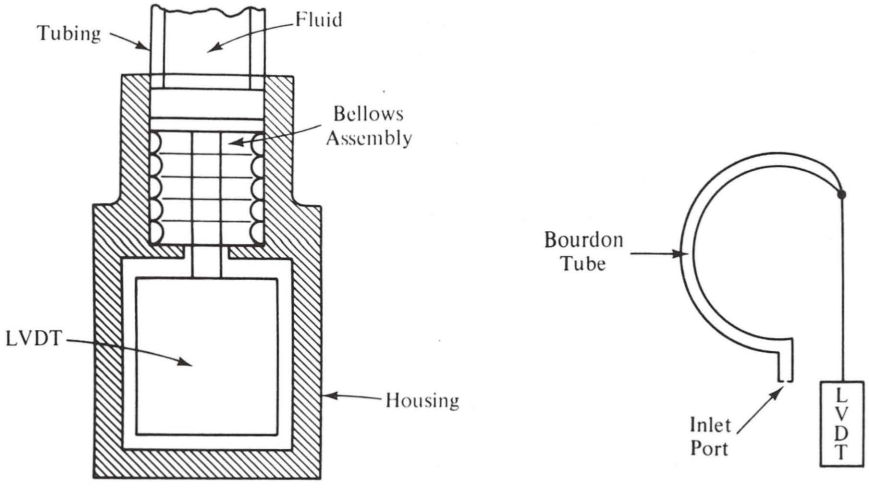
13-20 FLUID PRESSURE TRANSDUCERS

Fluid pressures are measured in a variety of ways, but the most common involve a transducer such as those shown in Figs. 13-18 and 13-19.

In the example of Fig. 13-18(a) a strain gage or LVDT is mounted inside a housing that has a bellows or aneroid assembly exposed to the fluid. More force is applied to the LVDT or gage assembly as the bellows compresses. The compression of the bellows is proportional to the fluid pressure.

An example of the *Bourdon tube* pressure transducer is shown in Fig. 13-18(b). Such a tube is hollow and curved, but flexible. When a pressure is applied through the inlet port, the tube tends to straighten out. If the end tip is connected to a position/displacement transducer, then the transducer output will be proportional to the applied pressure.

Figure 13-19(a) shows another popular form of fluid transducer. In this version, a diaphragm is mounted on a cylindrical support similar to Fig. 13-18. In some cases, a bonded strain gage is attached to the under side of the diaphragm, or flexible supports to an unbonded type are used. In the example shown, the diaphragm is connected to the core drive bar of an inductive transducer or LVDT.



(a) (b)

Figure 13-18. (a) Fluid pressure transducer. (b) Bourdon tube fluid pressure transducer.

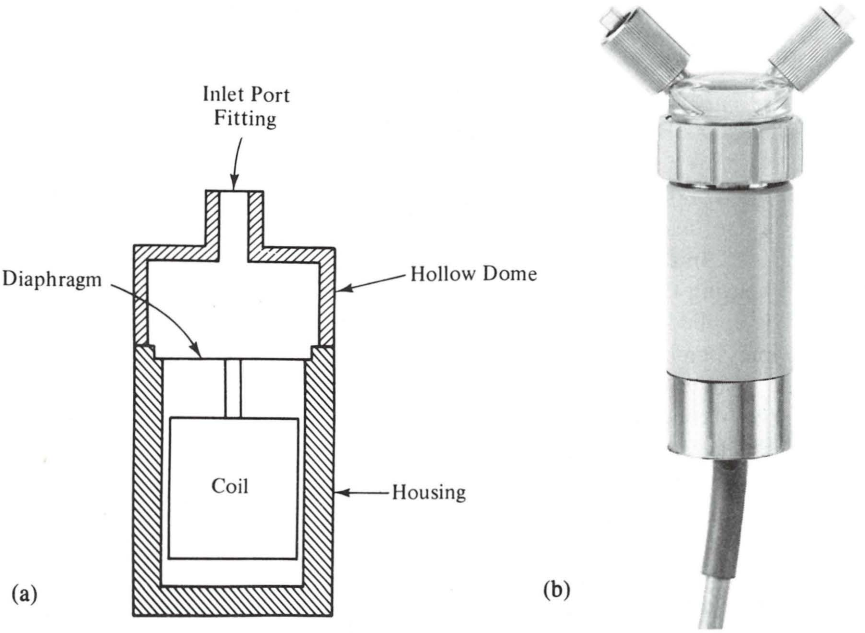


Figure 13-19. (a) Dome-type fluid pressure transducer. (b) Commercial dome-type transducer. (Courtesy of Hewlett-Packard.)

Figure 13-19(b) shows the Hewlett-Packard type 1280 transducer used in medical electronics to measure human blood pressure. In this device, the hollow fluid-filled dome is fitted with *Luer-lock* fittings, standards in medical apparatus.

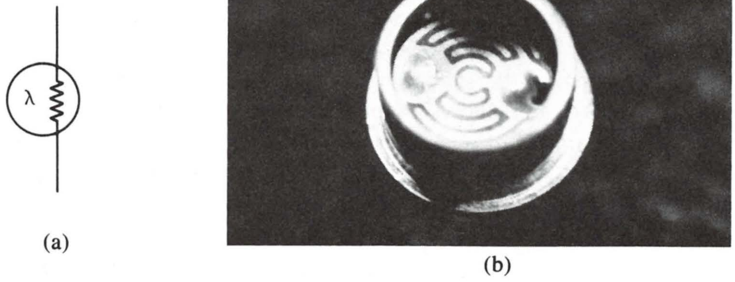
The fluid transducers shown so far will measure gage pressure, i.e., pressure above atmospheric pressure, because one side of the diaphragm is open to air. A *differential* pressure transducer will measure the difference between pressures applied to the two sides of the diaphragm. Such devices will have two ports marked such as "*P*1" and "*P*2," or something similar.

13-21 LIGHT TRANSDUCERS

There are several different phenomena for measuring light, and they create different types of transducer. For this chapter, we will limit the discussion to *photoresistors, photovoltaic cells, photodiodes,* and *phototransistors*.

A photoresistor can be made because certain semiconductor elements show a marked decrease in electrical resistance when exposed to light. Most materials do not change linearly with increased light intensity, but certain combinations such as cadmium sulphide (CdS) and cadmium selenide (CdSe) are effective. These cells operate over a spectrum from "near-infrared" through most of the visible light range, and can be made to operate at light levels of 10^{-3} to 10^{+3} footcandles (i.e., 10^{-3} to 70 mW/cm^2). Figure 13-20(a) shows the photoresistor circuit symbol, while Fig. 13-20(b) shows an example of a photoresistor.

A photovoltaic cell, or "solar cell," as it is sometimes called, will produce an electrical current when connected to a load. Both silicon (Si) and selenium (Se) types are known. The Si type covers the visible and near-infrared spectrum,



(a)

(b)

Figure 13-20. (a) Symbol for photoresistor cell. (b) Actual photoresistor cell.

at intensities between 10^{-3} and 10^{+3} mW/cm^2. The selenium cell, on the other hand, operates at intensities of 10^{-1} to 10^2 mW/cm^2, but accepts a spectrum of near-infrared to the ultraviolet.

Semiconductor pn junctions under sufficient illumination will respond to light. Interestingly enough, they tend to be photoconductive when heavily reverse-biased, and photovoltaic when forward-biased. These phenomena have led to a whole family of photodiodes and phototransistors.

13-22 CAPACITIVE TRANSDUCERS

A parallel plate capacitor can be made by positioning two conductive planes parallel to each other. The capacitance is given by

$$C = \frac{kKA}{d} \tag{13-25}$$

where C = the capacitance in *farads* (F), or a subunit (μF, pF, etc.)
k = a units constant
K = the dielectric constant of the material used in the space between the plates (K for air is 1)
A = the area of the plates "shading" each other
d = the distance between the plates

Figure 13-21 shows several forms of capacitance transducer. In Fig. 13-21(a) we see a rotary plate capacitor that is not unlike the variable capacitors used to tune radio transmitters and receivers. The capacitance of this unit is proportional to the amount of area on the fixed plate that is covered, i.e., "shaded," by the moving plate. This type of transducer will give signals proportional to curvilinear displacement, or angular velocity.

A rectilinear capacitance transducer is shown in Fig. 13-21(b), and it consists of a fixed cylinder and a moving cylinder. These pieces are configured so that the moving piece fits inside of the fixed piece, but is insulated from it.

Two types of capacitive transducer discussed so far vary capacitance by changing the shaded area of two conductive surfaces. Figure 13-21(c), on the other hand, shows a transducer that varies the *spacing* between surfaces, i.e., the d term in Eq. (13-25). In this device, the metal surfaces are a fixed plate and a thin diaphragm. The dielectric is either air or a vacuum. Such devices are often used as capacitance microphones.

Capacitance transducers can be used in several ways. One method is to use the varying capacitance to frequency-modulate an RF oscillator. This is the method employed with capacitance microphones [those built like Fig. 13-21(c), not the electrotet type]. Another method is to use the capacitance transducer in an ac bridge circuit (Chapter 3).

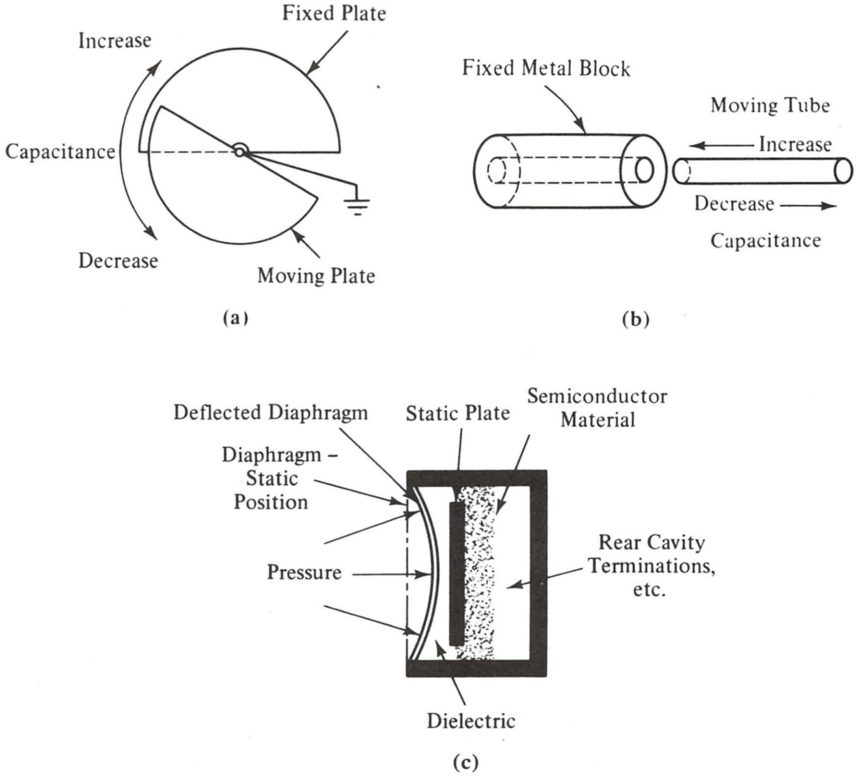


Figure 13-21. Capacitance transducers. (Source: Harry Thomas, *Handbook of Biomedical Instrumentation and Measurement* Figure 1-6, p. 13, Reston Publishing Co.)

Figure 13-22 shows an application in which a differential capacitance transducer is used to provide a position signal in a PMMC galvanometer chart recorder (Chapter 10). This type of transducer is superior to others in this application, because it can be constructed with very low mass, so will not significantly dampen the pen motion. The position signal is used in a feedback control circuit that gives the pen controlled damping. In one model, the differentiated position signal warns the instrument that the pen will slam into the mechanical limit stop, and puts the brakes on to prevent pen damage.

13-23 SUMMARY

1. A *transducer* is a device that converts a physical stimulus such as displacement, force, temperature, etc., to an electrical voltage or current proportional to the magnitude of the stimulus.

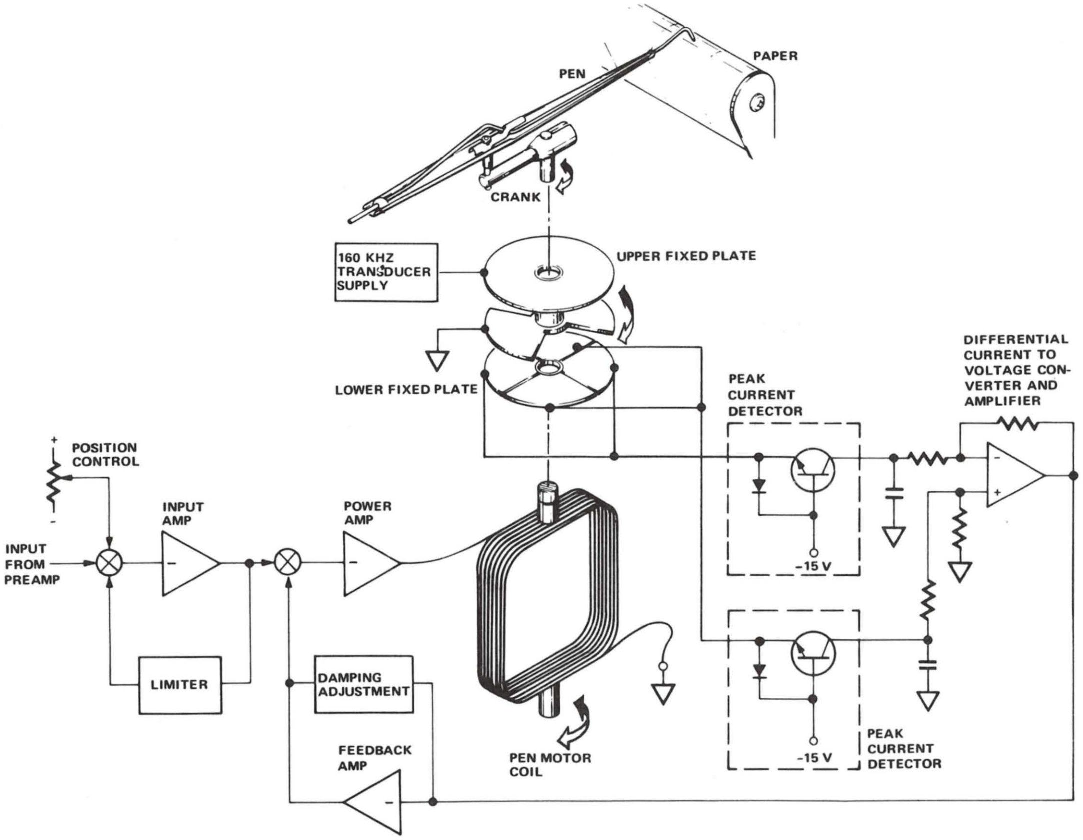


Figure 13-22. Capacitance transducers used as a position/velocity sensor. (Courtesy of Hewlett-Packard.)

2. A *strain gage* depends upon the *piezoresistance*—i.e., resistance change due to mechanical deformation—phenomenon.
3. There are two basic forms of inductive transducer: X_L bridge and LVDT.
4. Capacitive transducers depend upon varying the space between, or the shaded area of, two conductive surfaces.
5. Several types of temperature transducer are common: thermistor, thermocouple, and semiconductor junctions.

13-24 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

13-25 QUESTIONS

1. Define in your own words "transducer."
2. List several electrical parameters that can be used to transduce physical parameters.
3. Many transducers are often used in ____________ bridge circuits.
4. Define "piezoresistivity" in your own words.
5. *Tension* forces in a cylindrical conductor ____________ the electrical resistance.
6. *Compression* forces in a cylindrical conductor ____________ the electrical resistance.
7. Changes in electrical resistance due to deformation are a phenomenon called ____________.
8. Define in your own words "gage factor."
9. Write the two equations for gage factor.
10. Discuss the differences between *bonded* and *unbonded* strain gages.
11. Write the equation for E_0 in a Wheatstone bridge transducer that uses *one* active element.
12. What are the *units* for the sensitivity of the fluid pressure transducer?
13. Draw a circuit showing a *balance* control in a Wheatstone bridge strain gage.
14. List three types of temperature transducer.
15. Define *Seebeck effect* and *Peltier effect* in your own words. How are these effects related?
16. The temperature coefficients for most elemental metals are ____________.

17. Specially "doped" semiconductor materials can have both ____________ and ____________ temperature coefficients.
18. A thermistor used as an electronic thermometer must be operated to minimize ____________.
19. A thermocouple is formed of two ____________ metals, and works because the metals have different ____________ ____________.
20. List three types of semiconductor temperature transducer.
21. List two types of inductive transducer. Draw appropriate circuit diagrams.
22. The secondary windings of an LVDT are connected in ____________.
23. List two types of position/displacement transducers.
24. Draw circuits for one- and two-quadrant position transducers.
25. How can a displacement transducer derive velocity and acceleration signals?
26. What types of device are used to measure angular velocity?
27. List three types of phototransducer.
28. List two types of photoconductive semiconductor element.

13-26 PROBLEMS

1. A constantan cylinder with a diameter of 0.05 mm is 14.6 cm long. Find its electrical resistance.
2. Predict the *gage factor* if a 12-mm-long strain gage element changes length 0.7 mm, and its resistance changes from 78 to 82 ohms. What type of force is applied, *tension* or *compression?*
3. What is the *gage factor* of a cylindrical wire 0.05 mm in diameter and 18 mm long if the diameter changes 0.03 mm and the length changes 2.9 mm when a tension force is applied?
4. A Wheatstone bridge strain gage is excited by a 6-volt dc source. When it is stimulated by an outside parameter each element changes resistance by a factor of $0.042R$, and all four arms have equal resistance under zero-stimulus conditions. Calculate the output voltage E_0 for (a) one, (b) two, and (c) three active elements.
5. A 5-gram mass hangs from the end of a bar-type strain gage (i.e., Fig. 13-12). Express this mass as a force in *dynes*.
6. A fluid pressure transducer has a sensitivity of 50 μV/V/cm Hg and is excited by a +7.5-volt dc source. Find the output voltage if a 150-mm Hg pressure is applied.
7. The transducer in Problem 6 is a Wheatstone bridge strain gage, with four active elements that each have a zero-pressure resistance of 500 ohms. Calculate the value of a calibration resistance (to shunt one arm) that will give an artificial pressure of 120 mm Hg.

8. A thermistor is known to have a resistance of 100 kΩ at 25°C, and a temperature coefficient of $0.02°\mathrm{C}^{-1}$. Calculate the resistance at (a) 30°C and (b) 56°C.
9. Find the resistance of a thermistor at 100°C if the ice-point resistance is 500 kΩ, and the value of β is 2900°K.
10. Find the base-emitter voltage of a transistor at 33°C if the collector current is 10 mA, and the reverse saturation current is 10^{-13} ampere.
11. Find the slope factor of a dual-transistor temperature transducer such as Fig. 13-9 if $I1 = 10$ mA and $I2 = 3$ mA.
12. What amplification following the transducer in Problem 11 results in a scale factor of 10 mV/°K?

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chapter 14

DATA CONVERTERS

14-1 OBJECTIVES

1. To learn the operation of digital-to-analog converters (DACs).
2. To learn the operation of analog-to-digital converters (ADCs).
3. To learn the principal applications of data converters.
4. To learn the key design parameters used to select data converters.

14-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. Most binary DAC circuits use either the ____________ ____________ or ____________ resistance ladders.
2. List three types of ADC circuit.
3. How many different *states* can be represented on an eight-bit DAC? What is the highest *number* that can be represented on an eight-bit DAC?
4. Each step in a 10-bit DAC with a full-scale output potential of 10.00 volts represents an output voltage change of ____________ mV.

14-3 WHAT ARE DATA CONVERTERS?

Analog circuits and digital instruments occupy mutually exclusive realms. In the analog world, a signal may vary between upper and low limits, and may assume any value within the range. Analog signals are continuous between limits. But the signals in digital circuits may assume only one of two discrete voltage levels, i.e., one each for the two binary digits, 0 and 1. A *data converter* is a circuit or device that examines signals from one of these realms and then converts it to a proportional signal from the other.

A *digital-to-analog converter* (DAC), for example, converts a digital (i.e., binary) "word" consisting of a certain number of *bits* into a *voltage* or *current* that represents the binary number value of the digital word. An 8-bit DAC, for example, may produce an output signal of 0 volts when the binary word applied to its digital inputs is 00000000_2, and (say), 2.56 volts when the digital inputs see a word of 11111111_2. For binary words applied to the inputs, then, a porportional output voltage is created.

In the case of an *analog-to-digital converter* (ADC, or A/D), an analog voltage or current produces a proportional binary word output. If an 8-bit ADC has a 0- to 2.56-volt input signal range, then 0 volt input could produce an output word of 00000000_2, while the +2.56-volt level seen at the input would produce an output word of 11111111_2.

Data converters are used primarily to interface transducers (most of which produce analog output signals) to digital instruments or computer inputs, and to interface digital instrument outputs to analog-world devices such as meter movements, chart recorders, motors, etc.

We have already discussed one form of ADC circuit, the dual-slope integrator, in Chapter 6. It is the primary ADC used to make digital voltmeters. In this chapter we will consider common DAC circuits, and the following ADC circuits: *servo* (also called the *binary counter* or *ramp* ADC), *successive approximation, parallel converter*, and the *voltage-to-frequency converter*.

14-4 DAC CIRCUITS

Figure 14-1 shows a *binary weighted resistance ladder* and operational amplifier used as a binary DAC. The operation of this circuit can be deduced from operational amplifier theory given in Chapter 12.

The resistors in the ladder are said to be binary weighted, because their values are related to each other by powers of two. If the lowest-value resistor is given the value R, then the next in the sequence will have a value $2R$, followed by $4R$, $8R$, $16R$, all the way up to the nth resistor (last one in the chain), which has a value of $(2^{(n-1)})R$.

The switches ($B1$ through B_n) represent the input bits of the digital word. Although shown here as mechanical switches, they would be transistor switches in actual practice. The switches are used to connect the input resistors to either ground, or voltage source E, to represent binary states 0 and 1, respectively. Switches $B1$ through B_n create currents $I1$ through I_n, respectively, when they are set to the "1" position.

We know from Ohm's law that each current $I1$ through I_n is equal to the quotient of E and the value of the associated resistor, i.e.:

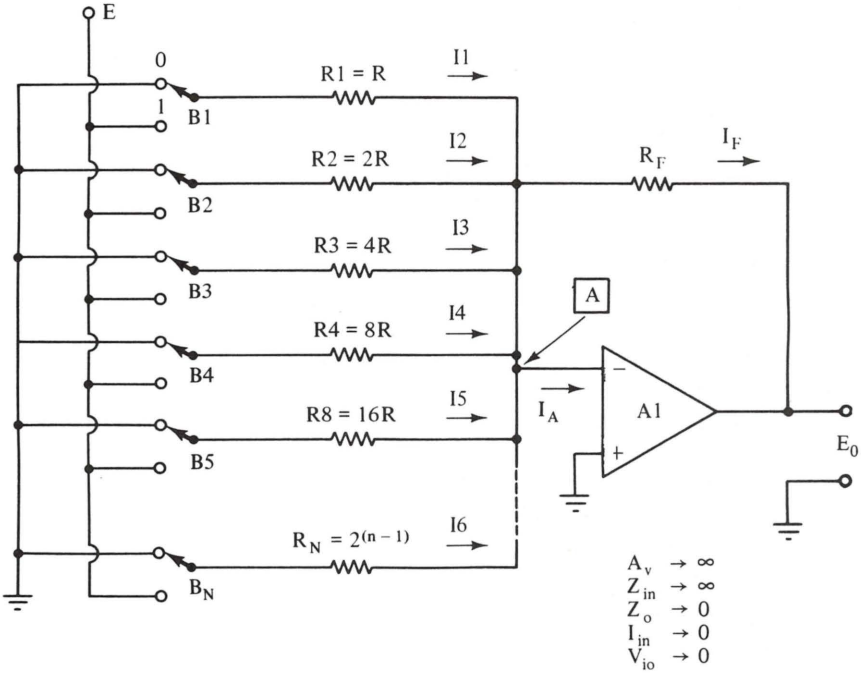


Figure 14-1. Binary weighted resistor ladder DAC circuit.

$$
\begin{array}{lclcl}
I1 & = & E/R1 & = & E/R \\
I2 & = & E/R2 & = & E/2R \\
I3 & = & E/R3 & = & E/4R \\
\bullet & & \bullet & & \bullet \\
\bullet & & \bullet & & \bullet \\
\bullet & & \bullet & & \bullet \\
I_n & = & E/R_n & & E/(2^{(n-1)})R
\end{array}
$$

The total current into the junction (point A in Fig. 14-1) is expressed by the summation of currents $I1$ through I_n:

$$
I_A = \sum_{i=1}^{n} \frac{a_i E}{2^{(i-1)} R} \tag{14-1}
$$

where I_A = the current into the junction (point A) in amperes (A)
E = the reference potential in volts (V)
R = the resistance of $R1$ in ohms (Ω)

a_i = either "1" or "0," depending upon whether the input bit is "1" or "0"

n = the number of bits, i.e., the number of switches

From operational amplifier theory we know that

$$I_A = -I_f \tag{14-2}$$

and

$$E_0 = I_f R_f \tag{14-3}$$

So, by substituting Eq. (14-2) into Eq. (14-3), we obtain

$$E_0 = -I_A R_f \tag{14-4}$$

and substituting Eq. (14-1) into Eq. (14-4) yields

$$E_0 = -R_f \sum_{i=1}^{n} \frac{a_i E}{2^{(i-1)} R} \tag{14-5}$$

Since E and R are constants, we usually write Eq. (14-5) in the form

$$E_0 = \frac{-ER_f}{R} \sum \frac{a_i}{2^{(i-1)} R} \tag{14-6}$$

Example 14-1

A four-bit (i.e., $n = 4$) DAC using a binary weighted resistor ladder has a reference source of 10 volts dc, and $R_f = R$. Find the output voltage E_0 for the input word 1011_2. (*Hint:* For input 1011, $a_1 = 1$, $a_2 = 0$, $a_3 = 1$, and $a_4 = 1$.)

SOLUTION

$$E_0 = \frac{-ER_f}{R} \sum \frac{a_i}{2^{(i-1)} R} \tag{14-6}$$

$$E_0 = \left[\frac{(-10\text{ V})(R)}{(R)}\right] \left[\frac{1}{2^{(1-1)}} + \frac{0}{2^{(2-1)}} + \frac{1}{2^{(3-1)}} + \frac{1}{2^{(4-1)}}\right]$$

$$E_0 = (-10\text{ V})\left[\frac{1}{2^0} + \frac{1}{2^2} + \frac{1}{2^3}\right]$$

$$E_0 = (-10\text{ V})\left[\frac{1}{1} + \frac{1}{4} + \frac{1}{8}\right]$$

$$E_0 = (-10\text{ V})\left[\frac{1}{1.375}\right] = \mathbf{-7.27\ volts}$$

Although not revealed by the idealized equations, the binary weighted resistance ladder suffers from a serious drawback in actual practice. The values of the input resistors tend to become very large and very small at the ends of the range as the bit length of the input word becomes longer. If R is set to 10 kΩ (a popular value), then $R8$ will be 1.28 megohms. If we assume a reference potential E of 10.00 volts dc, then $I8$ will be only 7.8 *microamperes*. Most common nonpremium-grade operational amplifiers will not be able to resolve signals that low from the inherent noise. As a result, the bit length of the binary weighted ladder is severely limited. Few of these types of converters are found with more than 6- or 8-bit word lengths.

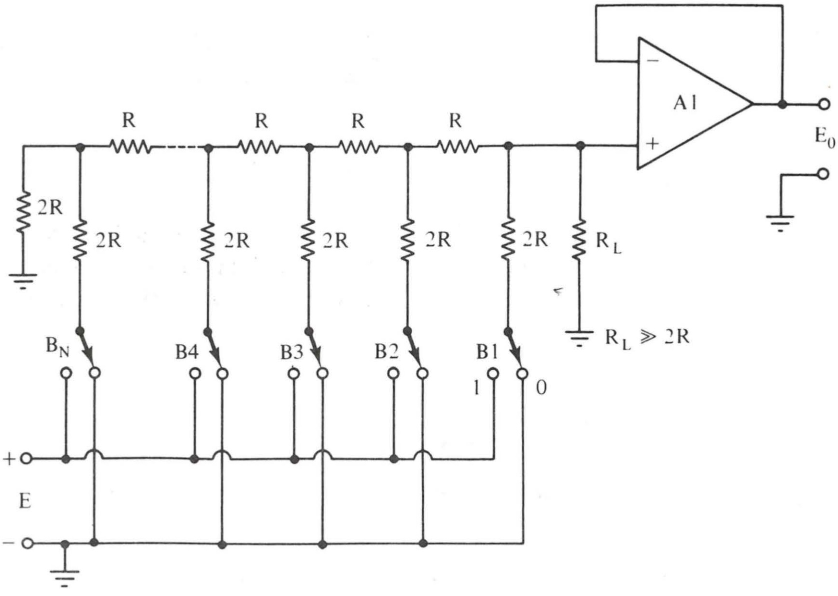


Figure 14-2. R-2R resistor ladder DAC circuit.

In commercial DACs, all of the resistors have a value of either R or $2R$. The gain of the amplifier is unity, so E_0 can be expressed as

$$E_0 = (E) \sum_{i=1}^{n} \frac{a_i}{2_i} \tag{14-7}$$

(provided that $R_L >> R$, so that the voltage divider effect between the ladder and R_L can safely be neglected).

Example 14-2

A four-bit DAC using the R-$2R$ technique has a 5.00-volt dc reference potential. Calculate E_0 for the input word 1011_2.

SOLUTION

$$E_0 = E \sum_{i=1}^{n} \frac{a_i}{2^i} \tag{14-7}$$

$$E_0 = (5\ \text{V}) \left[\frac{1}{2^1} + \frac{0}{2^2} + \frac{1}{2^3} + \frac{1}{2^4} \right]$$

$$E_0 = (5\ \text{V}) \left[\frac{1}{2} + 0 + \frac{1}{8} + \frac{1}{16} \right]$$

$$E_0 = (5\ \text{V})(0.688) = \mathbf{3.44\ volts}$$

The full-scale output voltage for any DAC using the R-$2R$ resistor ladder is given by

$$E_{fs} = \frac{E(2^n - 1)}{2^n} \tag{14-8}$$

where E_{fs} = the full-scale output potential in volts (V)
E = the reference potential in volts (V)
n = the bit length of the digital input word

Example 14-3

Find the full-scale output potential for an 8-bit DAC with a reference potential of 10.00 volts dc.

SOLUTION

$$E_{\mathrm{fs}} = \frac{E(2^{\mathrm{n}} - 1)}{2^{\mathrm{n}}} \qquad (14\text{-}8)$$

$$E_{\mathrm{fs}} = \frac{(10\ \mathrm{V})(2^8 - 1)}{2^8}$$

$$E_{\mathrm{fs}} = \frac{(10\ \mathrm{V})(2^7)}{(2^8)}$$

$$E_{\mathrm{fs}} = \frac{(10\ \mathrm{V})(255)}{(256)} = \mathbf{9.96\ volts}$$

The output of a DAC cannot change in a continuous manner, because the input is a digital word; i.e., it can exist only in certain discrete states. Each successive binary number changes the output an amount equal to the change created by the least significant bit (LSB), which is expressed by

$$\Delta E_0 = \frac{E}{2^{\mathrm{n}}} \qquad (14\text{-}9)$$

So, for the DAC in Example 14-3, E_0 would be

$$\Delta E_0 = \frac{(10\ \mathrm{V})}{2^8}$$

$$\Delta E_0 = \frac{(10\ \mathrm{V})}{(256)} = \mathbf{40\ mV}$$

ΔE_0 is often called the 1 LSB value of E_0, and is the smallest change in output voltage that can occur. It is interesting that, if we let 0 volts represent 00000000_2 in our 8-bit system, then the maximum value of E_0 at 11111111_2 will be 1 LSB less than E (confirmed by the result of Example 14-3).

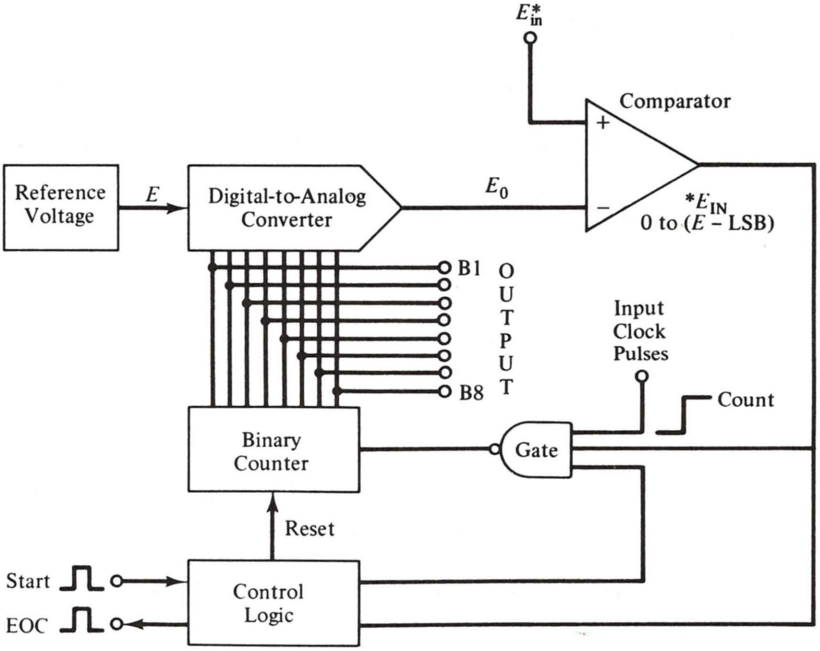
There are numerous commercial DACs on the market in IC, function module block, and equipment form. The reader should consult manufacturer's catalogues for appropriate types in any given application.

14-5 SERVO ADC CIRCUITS

The *servo* ADC circuit (also called *binary counter* or *ramp* ADC circuit) uses a binary counter to drive the digital inputs of a DAC. A voltage comparator keeps the clock gate to the counter open as long as $E_0 \neq E_{\text{in}}$.

An example of such a circuit is the 8-bit ADC in Fig. 14-3(a), while the relationship of E_0 and E_{in} relative to time is shown in Fig. 14-3(b).

Two things happen when a *start* pulse is received by the control logic circuits: the binary counter is reset to 00000000_2, and the gate is opened to allow clock pulses into the counter. This will permit the counter to begin incrementing, thereby causing the DAC output voltage E_0 to begin rising [Fig. 14-3(b)]. E_0 will



(a)

Figure 14-3. (a) Servo ADC circuit.

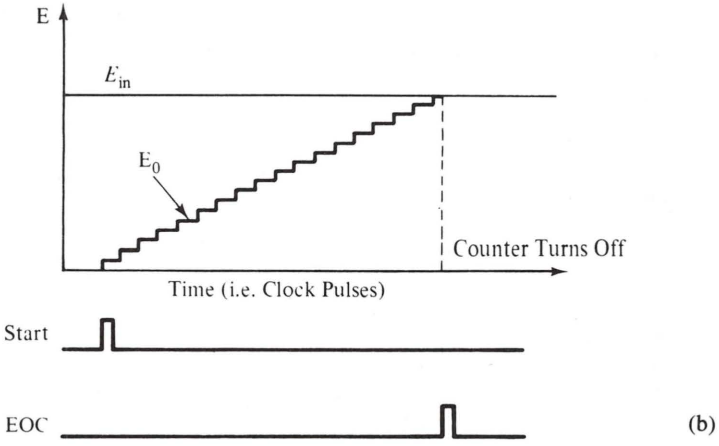


Figure 14-3. ***(cont.)*** (b) Operation of the servo-type ADC circuit.

continue to rise until $E_0 = E_{in}$. When this condition is met, the output of the comparator drops *low*, turning off the gate. The binary number appearing on the counter output at this time is proportional to E_{in}.

The control logic section senses the change in comparator output level, and uses it to issue an *end-of-conversion* (EOC) pulse. This EOC pulse is used by instruments or circuitry connected to the ADC to verify that the output data are valid.

The *conversion time* T_c of an ADC such as this depends upon the value of E_{in}, so when E_{in} is maximum (i.e., full scale), so is T_c. Conversion time for this type of ADC is on the order of 2^n clock pulses for a full-scale conversion.

14-6 SUCCESSIVE APPROXIMATION ADC CIRCUITS

The conversion time of the servo ADC is too long for some applications. The successive approximation (SA) ADC is much faster for the same clock speed; i.e., it takes $(n + 1)$ clock pulses instead of 2^n. For the 8-bit ADC that has been our example, the SA type of ADC is 28 times *faster* than the *servo* ADC.

The basic concept of the SA ADC circuit can be represented by a platform balance, such as Fig. 14-4, in which a full-scale weight W will deflect the pointer all the way to the left when pan 2 is empty.

Our calibrated weight set consists of many separate pieces, which weigh $W/2$, $W/4$, $W/8$, $W/16$, etc. When an unknown weight W_x is placed on pan 2, the scale will deflect to the right. To make our measurement, we start with $W/2$, and place it on pan 1. Three conditions are now possible:

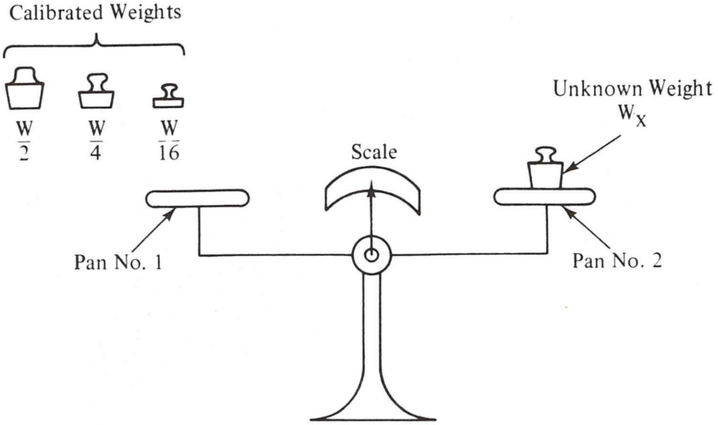


Figure 14-4. Successive approximation ADC circuits are like the platform balance.

$W/2 = W_X$ (scale is at zero)
$W/2 > W_X$ (scale is to the left of zero)
$W/2 < W_X$ (scale is to the right of zero)

If $W/2 = W_X$, then the measurement is finished, and no additional trials are necessary. But if $W/2$ is less than W_X, then we must add more weights in succession ($W/4$, then $W/8$, etc.) until we find a combination equal to W_X.

If, on the other hand, $W/2$ is greater than W_X, then we must *remove* the $W/2$ weight, and in the second trial start again with $W/4$. This procedure will continue until a combination equal to W_X is found.

In the SA ADC circuit we do not use a scale, but a shift register, as in Fig. 14-5. A successive approximation register (SAR) contains the control logic, a shift register, and a set of output latches, one for each register section. The outputs of the latches drive a DAC.

A *start* pulse sets the first bit of the shift register *high,* so the DAC will see the word 10000000_2, and therefore produces an output voltage equal to one-half of the full-scale output voltage. If the input voltage is greater than ½E_{fs}, then the $B1$ latch is set *high.* On the next clock pulse, register $B2$ is set high for trial 2. The output of the DAC is now ¾-scale. If, on any trial, it is found that $E_{in} < E_0$, then that bit is reset *low*.

Let us follow a 3-bit SAR through a sample conversion. In our example, let us say that the full-scale potential is 1 volt, and E_{in} is 0.625 volt. Consider Fig. 14-6.

Time t_1: The *start* is received, so register $B1$ goes *high.* Output word is now 100_2, so $E_0 = 0.5$ V. Since E_0 is less than E_{in}, latch $B1$ is set to "1," so at the end of the trial, the output word remains 100_2.

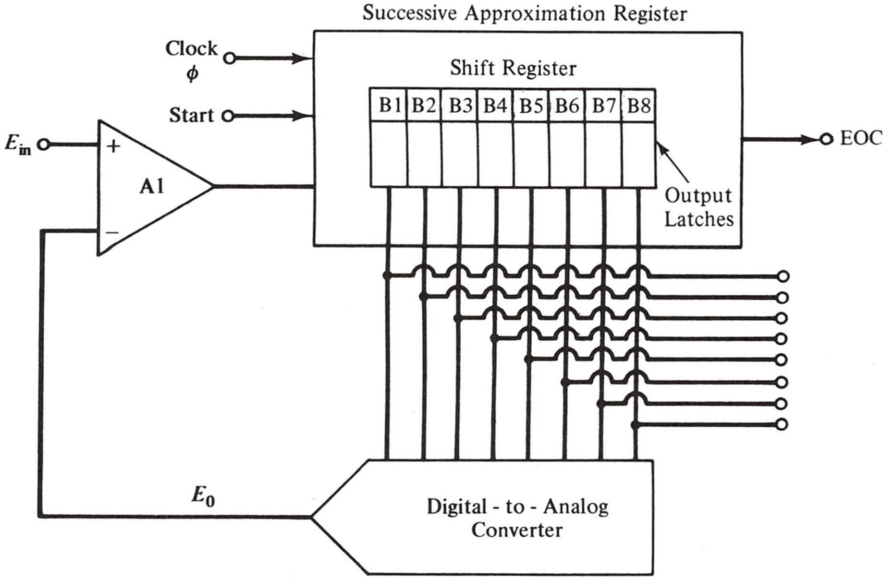


Figure 14-5. Successive approximation ADC circuit.

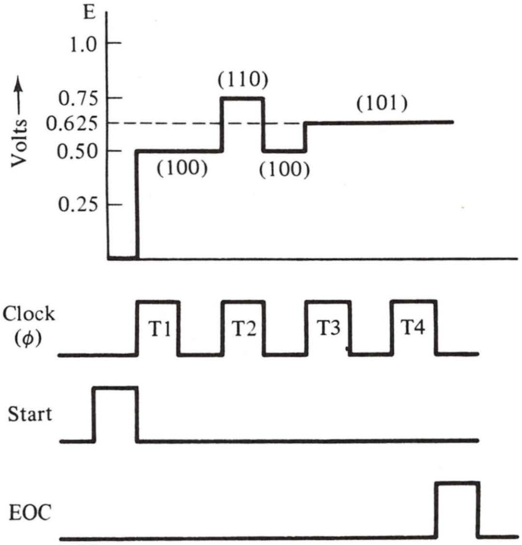


Figure 14-6. Timing diagram for Fig. 14-5.

Time t_2: On this trial (which starts upon receiving the next clock pulse), register $B2$ is set *high,* so the output word is 110_2. Voltage E_0 is now 0.75 volt. Since E_{in} is less than E_0, the $B2$ latch is set to "0," and the output word reverts to 100_2.

Time t_3: Register $B3$ is set *high,* making the output word 101_2. The value of E_0 is now 0.625 volt, so $E_{\text{in}} = E_0$. The $B3$ register is latched to 1, and the output word remains 101_2.

Time t_4: Overflow occurs, telling the control logic to issue an EOC pulse. In some cases the overflow pulse *is* the EOC pulse.

Note that in the example, we had a 3-bit SAR, so by our $(n + 1)$ rule, required four clock pulses to complete the conversion. The SA type of ADC was once regarded as difficult to design because of the logic required. But today, IC and function blocks are available that use this technique, so the design job is reduced considerably. The SA technique can be implemented in software under computer control using only an external DAC and comparator. All register functions are handled in the software (program). (See references 1, 2, and 3 on page 328.)

14-7 PARALLEL CONVERTERS

The parallel ADC circuit (Fig. 14-7) is probably the fastest type of ADC known. In fact, some texts call it the "flash" converter in testimony to its speed. It consists of a bank of $(2^n - 1)$ voltage comparators biased by reference potential E through a resistor network that keeps the individual comparators 1 LSB apart. Since the input voltage is applied to all of the comparators simultaneously, the speed of conversion is essentially the slewing speed of the *slowest* comparator in the bank, and the decoder propagation time (if logic is used). The decoder converts the output code to binary code, or possibly BCD in some cases.

14-8 VOLTAGE-TO-FREQUENCY CONVERTERS

A voltage-to-frequency (V/F) converter is a voltage-controlled oscillator (VCO) in which an input voltage E_{in} is represented by an output frequency F. An ADC using the V/F converter is shown in Fig. 14-8. It consists of no more than the VCO and a frequency counter. The display, or output states, on the counter gives us the value of E_{in}.

Voltage-to-frequency converters are used mainly where economics dictate serial transmission of data from a remote collection point to the instrument. Such data can be transmitted by wire or radio communications channels. Another ap-

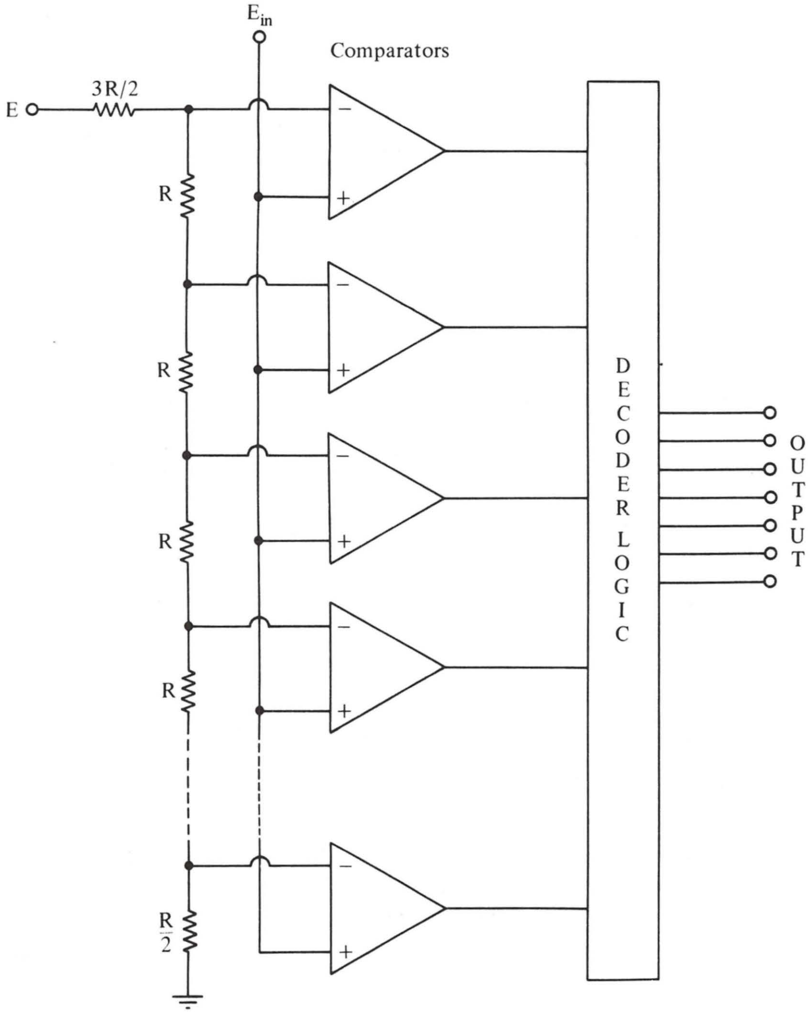


Figure 14-7. Parallel or "flash" ADC circuit.

plication is the tape recording of analog data that is, in itself, too low in frequency to be recorded.

The inverse procedure, F/V conversion, is a form of DAC, in which an input *frequency* is converted to an output *voltage*.

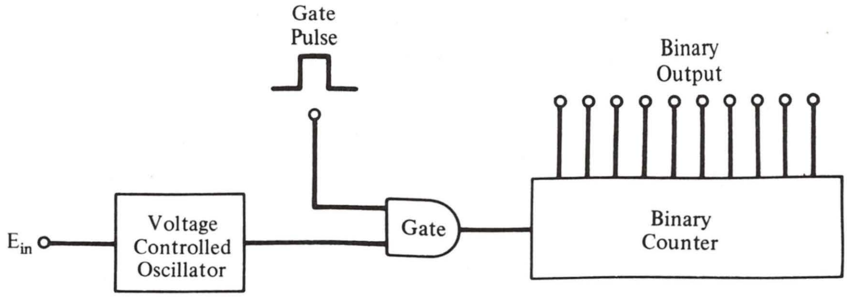


Figure 14-8. Voltage-to-frequency converter ADC.

14-9 SUMMARY

1. Analog-to-digital converters (ADC) convert analog voltages or currents to binary words.
2. Digital-to-analog converters (DAC) convert binary words to proportional voltages and currents.
3. Several types of ADC are in common use: dual-slope integrators, servo, successive approximation, and V/F converters.
4. There are two common forms of DAC, distinguished by the type of resistance ladder used: binary weighted or *R*-2*R*.

14-10 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

14-11 QUESTIONS

1. Define in your own words the purposes of (a) the DAC and (b) the ADC.
2. Draw a circuit for a binary weighted resistor ladder DAC.
3. Draw a circuit for an *R*-2*R* resistor ladder DAC.
4. List two types of ADC that use a DAC as a critical element.
5. The output of a DAC is a ____________ function of voltage or current.
6. Describe in your own words the operation of a successive approximation ADC.

7. Describe in your own words the operation of a servo ADC.
8. A very fast ADC circuit that uses a bank of voltage comparators is called the ____________ or ____________ ADC.

14-12 PROBLEMS

1. An 8-bit DAC using a binary weighted resistor ladder has a +7.5-volt dc reference potential. Find the output potential for the input word 10110111_2 if $R_f = R$.
2. An 8-bit DAC using an R-$2R$ ladder has a +2.56-volt dc reference potential. Find the full-scale output potential.
3. Find the 1 LSB output voltage for the DAC in Problem 2.
4. Find the conversion time for a full-scale input potential for (a) servo ADC, and (b) SA ADC, if the clock speed is 2.5 mHz.

14-13 REFERENCES

1. Don Aldridge, "A/D Conversion Systems: Let your μP do the Working," *EDN Magazine* (May 5, 1976), p. 75.
2. Wes Freeman and Will Ritmanich, "Cut A/D Conversion Costs by Using D/A Converters and Software," *Electronic Design* (April 26, 1977), p. 86.
3. "Software Controlled Analog to Digital Conversion Using the DAC-08 and the 8080A Microprocessor," Precision Monolithics, Inc., Application note AN-22, 1976.
4. Daniel H. Sheingold, *Analog-to-Digital Conversion Handbook,* Analog Devices, Inc. (Norwood, Mass).

chapter 15

PROBES, CONNECTORS, ETC.

15-1 OBJECTIVES

1. To learn about the types of connectors and probes used with instrumentation.
2. To learn some of the problems associated with probes.
3. To learn which probe or connector is appropriate in certain measurements.
4. To learn the correct *use* of probes.

15-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject matter. If you cannot answer any particular question, then look for the answer as you read the text.

1. Draw the circuit for a low-capacitance probe.
2. What is the *purpose* of a low-capacitance probe?
3. Why does a "twisted pair" sometimes perform better than test leads?
4. Why is it improper to use the power line grounding instead of direct probe grounding?

15-3 WHAT ARE "PROBES"?

A *probe,* as used in the context of electronic instrumentation, is a device used to connect the input of a measurement instrument such as an oscilloscope or electronic voltmeter to a point in the circuit where the measurement is to be made.

The probe may consist of alligator clip leads, or it may be a complex circuit including amplifiers and other active components. Along with probes, we will consider certain other hardware that is frequently used to interconnect instruments. We will also discuss when and where certain types of interconnecting device are suitable.

15-4 TEST LEADS

A test lead may be a length of hook-up wire with alligator clips at either end, or it may have a special probe-type end (Fig. 15-1). The probe usually makes it easier to access points in the circuit, and is generally safer to use in live circuits than are alligator clips. Besides the electrical shock hazard, there is always the possibility of accidentally slipping off the test point and damaging the circuit. Solid-state components are notoriously unforgiving of such "goofs."

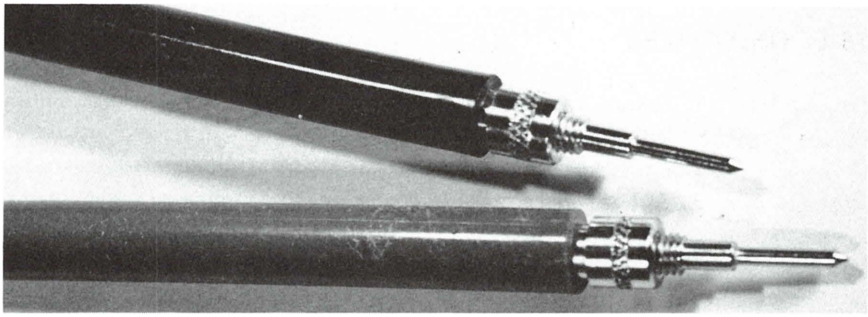


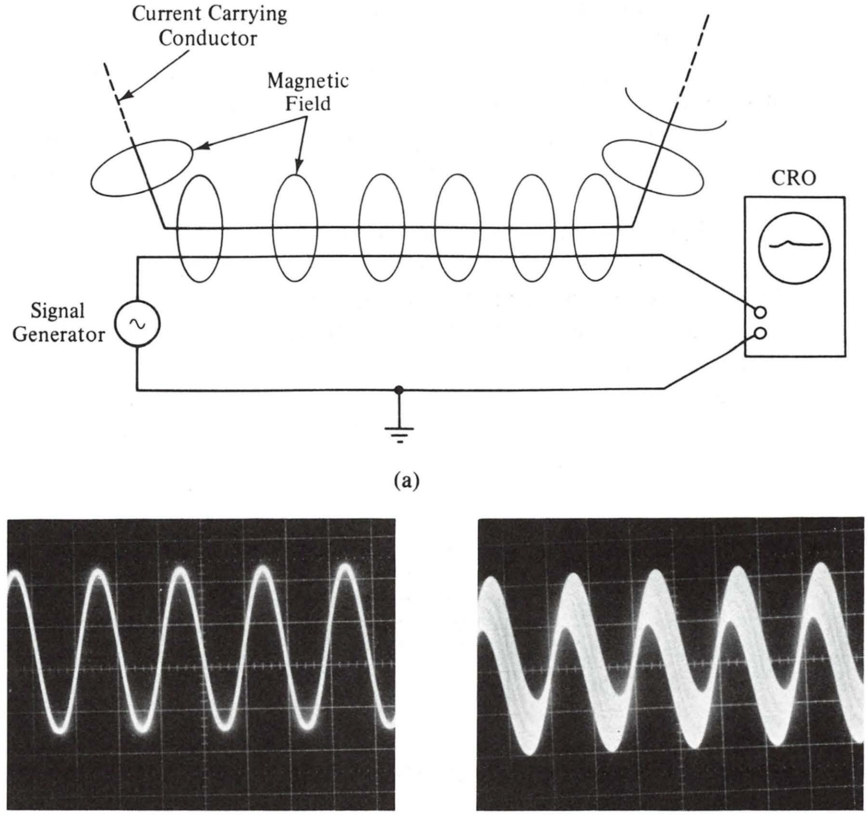
Figure 15-1. Test probes.

Test lead wire is a special type of hook-up wire that is well insulated (500–600 volts or more), but is flexible enough to allow easy use in making measurements.

The simple test lead is found most frequently in ac and dc voltmeter/multimeter applications, and in certain other applications where the ac frequency is low. At higher frequencies the distributed capacitance and noise pickup are too great for accurate measurement.

Figure 15-2(a) shows a mechanism by which interference can occur when open test leads are used. Note that one of the test lead wires is grounded, while the other carries a signal from a generator. A nearby conductor carrying an electric current generates a magnetic field, and flux from this field cuts the signal wire, inducing a spurious signal current. The "nearby" wire need not actually be in extremely close proximity to the signal wire, but is often the building power wiring located a dozen or so feet away from the test site. Indeed, the high current loads usually encountered in building power distribution systems render them quite capable of generating interference over a distance.

Figures 15-2(b) and 15-2(c) show the effect of building wiring originated 60-Hz interference on a 100-kHz signal circuit such as Fig. 15-2(a). The trace on the oscilloscope in Fig. 15-2(b) shows the 100-kHz, 100-mV signal *as it should be*. But in Fig. 15-2(c) we have the same trace modulated by the 60-Hz signal.



(a)

(b)

(c)

Figure 15-2. (a) Pickup of interference on unshielded test leads. (b) Proper reproduction of 100-kHz signal. (c) Reproduction with 60-Hz interference.

Note well that this effect is very often far more severe. It is possible to obtain up to several volts of 60-Hz signal in high impedance circuits, as demonstrated by touching the tip of an unterminated oscilloscope probe.

Besides frequency, two other factors must be considered when one is using test leads: *signal amplitude* and *circuit impedance*. If the amplitude is high enough, then the artifact caused by an interfering signal may be negligibly small. But the source and load impedances in the circuit affect the amplitude of the interfering signal. We observe several volts displayed on the screen of an oscilloscope, because the impedance of the connection between the probe tip and dry skin is in the range 10 kΩ to 1 megohm, and the oscilloscope input impedance is typically 1 megohm. In the example of Fig. 15-2 the signal generator supplying the 100-kHz signal has an output impedance of 600 ohms, so the circuit impedance is much lower. A much smaller amplitude signal, therefore, is displayed on the oscilloscope screen.

Note that a *current* is induced by the magnetic field, so by Ohm's law, if the impedance is high, then so is the voltage developed (and vice versa). There is a popular rule of thumb regarding 60-Hz pickup that claims 10 mV/in. but this figure is not dependable because of the dependence of the voltage on circuit impedances.

A possible solution that often proves effective is to use a *twisted pair* of test leads, as shown in Fig. 15-3. Such test leads can be purchased, or they can be made by inserting one end of each of the two wires into the chuck of an electric drill, and then anchoring the other ends. Some practice, however, is required before this job can be performed safely.

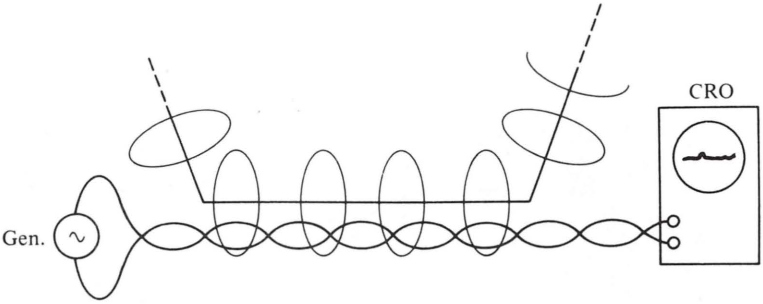


Figure 15-3. Twisted pair unshielded test leads.

The reasons for the effectiveness of the twisted pair are (1) self-shielding, and (2) the magnetic field affects both conductors equally. The self-shielding effect is particularly noted in unbalanced circuits where one side of the circuit is grounded, while the second reason is more prevalent where the two conductors are balanced with respect to ground.

The criteria for using simple test leads, then, are low frequency or dc signals, high signal amplitude, and low source and load impedances.

15-5 SHIELDED CABLES

One other possible solution is to use *shielded wire,* as shown in Fig. 15-4. A shielded wire is a conductor that is surrounded by and insulated from another conductor; in most cases the outer conductor is a braided cylinder. Coaxial cable, often used as a transmission line in RF circuits (Chapter 20), meets this description and is often used as shielded wire. Note, however, that shielded cables are

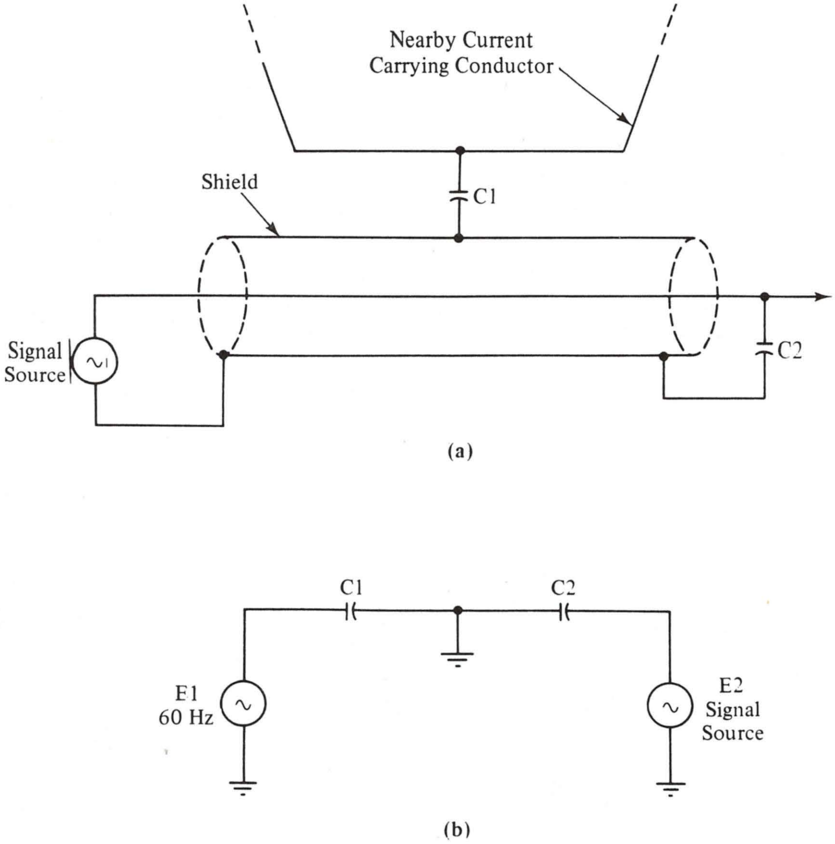


Figure 15-4. (a) Shielded test lead. (b) Equivalent circuit.

not necessarily transmission lines unless certain other criteria are met, and that includes coaxial cables.

The coupling between close-proximity conductors is both inductive and capacitive. Capacitor $C1$ in Fig. 15-4 is the capacitance between the shield and a nearby current-carrying conductor, and represents a path for interfering currents. Capacitor $C2$ in Fig. 15-4 represents the capacitance between the inner conductor and the shield.

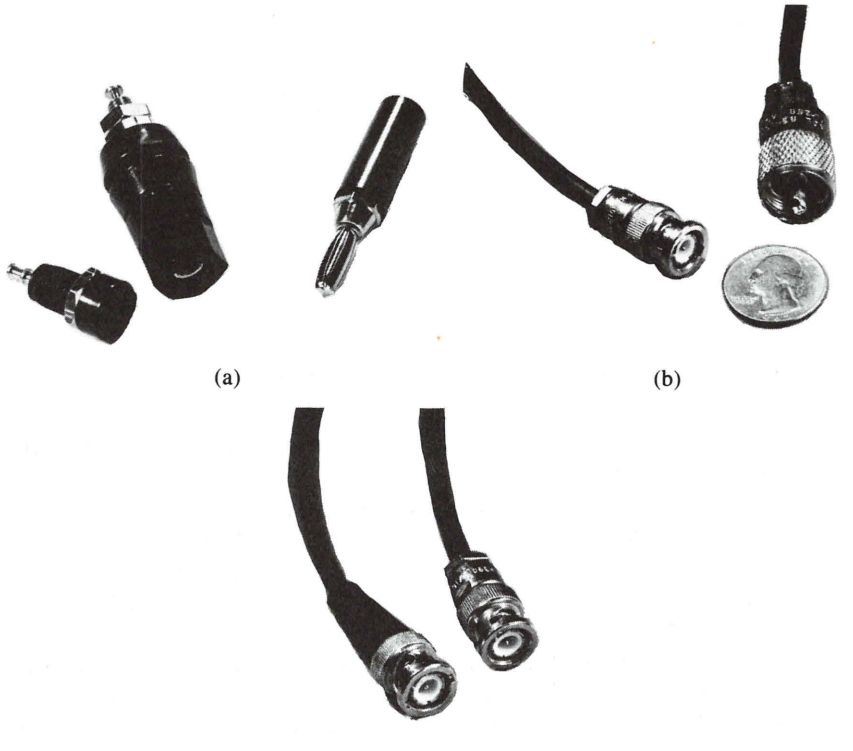
There are two main reasons why shielding can reduce coupling. In some systems, it is due to the fact that $C1$ and $C2$ are in *series,* so the total capacitance is less than either capacitance alone. In unbalanced systems such as Fig. 15-4 we find that one end of each capacitor is grounded, i.e., at zero potential. See Fig. 15-4(b). The interfering potential does not *usually* get into the signal circuit. For exceptions see Sec. 10-9.

15-6 CONNECTORS

There are quite a few different connectors used to interconnect electronic instrumentation and circuits, and (it seems to beleagured engineers and technicians) that the probability that two instruments will have different connectors is directly proportional to the urgency of using them together. All electronic facilities seem to keep a rather large number of adapters that are used to allow interfacing instruments that have different input or output connectors.

Figure 15-5 shows several different connectors. Figure 15-5(a) shows the banana plug and jack. These connectors are used on power supplies, audio generators, and other low-frequency instruments including multimeters and low-frequency oscilloscopes.

The PL-259 "UHF" connector is shown in Fig. 15-5(b). This connector mates with the SO-239 connector that is used as the antenna connector in CB and



(a)

(b)

(c)

Figure 15-5. (a) Banana plug and two types of banana jack. (b) BNC (*l*) and UHF (*r*) connectors. (c) Two types of BNC connector.

most other two-way radio transmitters. These connectors were once used extensively on signal generators and oscilloscopes built a decade or more in the past. Modern equipment, however, tends to use the BNC connector of Fig. 15-5(c). In current practice, the use of the UHF connector is limited to RF power applications below 450 mHz. Most owners of older instruments that were fitted with the SO-239 mate to the PL-259 have long since obtained a BNC-to-UHF adapter so that modern cables and probes will fit. A collection of adapters is shown in Fig. 15-6, although there are literally dozens on the market.

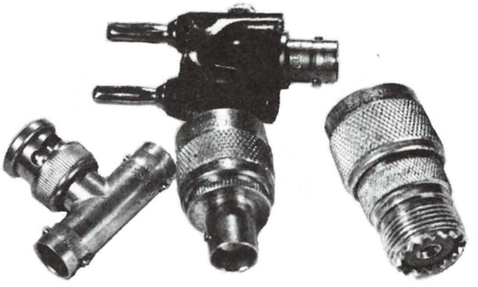


Figure 15-6. Connector adapters.

15-7 PROBLEMS WITH SHIELDED CABLES

Besides the ground-loop problems discussed in Sec. 10-9, we must also be on guard against the capacitance of shielded cables. That very property of the shielded cable that confers protection against one type of measurement disaster will provide equal opportunity for another. In some cases, the capacitance of the cable will severely attenuate high-frequency signals.

Although the figures vary from one type of cable to another, most small-diameter coaxial cables have a capacitance of approximately 30 pF/ft. A three-foot length of cable, then, has a capacitance of approximately 90 pF. Moreover, the input impedance of instruments such as oscilloscopes is usually specified in the parallel-equivalent form, i.e., a resistance shunted by a capacitance. Typical values are 1 megohm shunted by 30 pF. The source, then, sees a total impedance equivalent to 1 megohm shunted by 120 pF.

The source is not perfect either; it typically has a certain output impedance, so may be represented as in Fig. 15-7(a) by a resistor R_s in series with an ideal voltage source E (this E is very nearly the open-circuited output potential of the instrument). Figure 15-7(a) is an equivalent circuit that takes into account the various imperfections. Capacitors $C1$ and $C2$ represent the cable and instrument input capacitances, respectively; together they total 120 pF. Resistor $R1$ represents the oscilloscope input resistance. This circuit is essentially a voltage divider, and E_0 is the effective potential applied to the instrument input. To make

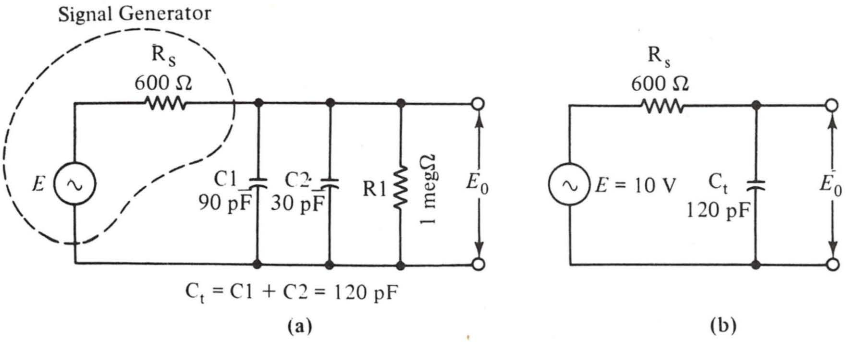


Figure 15-7. (a) Equivalent instrument input circuit with shielded leads. (b) Composite circuit.

the circuit arithmetic simpler, let us set E at 10 volts. At dc, the error that is due to voltage division is negligible:

$$E_0 = \frac{ER1}{R1 + R_s} \tag{15-1}$$

$$E_0 = \frac{(10\ \text{V})(1 \times 10^6\ \Omega)}{(1 \times 10^6 + 6 \times 10^2)\ \Omega}$$

$$E_0 = \frac{1 \times 10^6\ \text{V}}{1.0006 \times 10^6} = \mathbf{9.99\ V}$$

The error is approximately 0.1 percent. But consider what happens at higher frequencies, where the reactance of the capacitors can no longer be simply ignored. The reactance of a 120-pF capacitor at 100 kHz is approximately 13 kΩ, and at 1 mHz it drops to 1.3 kΩ. In both cases, the reactance of the capacitor would predominate, and the voltage output is reduced. We may ignore the 1-megohm input resistance, so our equivalent circuit is the RC voltage divider of Fig. 15-7(b). At 1 mHz the reactance of the capacitance is 1.3 kΩ, so E_0 will be

$$E_0 = \frac{(10\ \text{V})(1300\ \Omega)}{(1300 + 600)\ \Omega} \tag{15-2}$$

$$E_0 = \frac{(10\ \text{V})(1300)}{(1900)} = \mathbf{6.8\ V}$$

In addition to the voltage drop, there is a *phase shift* of the signal; i.e.,

$$\begin{aligned}\Theta &= \arctan (R_s/X_c) \qquad (15\text{-}3)\\ \Theta &= \arctan (600/1300)\\ \Theta &= \arctan (0.46) \approx \mathbf{28\ deg}\end{aligned}$$

These errors may not be terribly important on *some* sine wave measurements because a correction factor based on frequency can be applied, and phase shift may not be deemed important. But in making measurements to almost any nonsinusoidal waveform, however, these problems become acute. In fact, the faster the rise time of the waveform, the more acute they become. The solution to these problems is the use of a low-capacitance probe, described in Sec. 15-8.

Ordinary shielded cables should only be used for the same type of measurements as the test leads (Sec. 15-5). The shielded lead, however, usually confers the advantage over test leads of superior rejection of interference.

15-8 LOW-CAPACITANCE PROBES

The basic low-capacitance probe (Fig. 15-8) contains a parallel RC network consisting of a high-value resistor and a small-value capacitor. This type of probe is called a *passive* probe because it contains no amplifying devices. The probe in Fig. 15-8 presents one-tenth the capacitive load on the signal source as did the shielded cable, and provides a constant 10:1 voltage division ratio over a wide range of frequencies. Figure 15-9 shows a commercial low-capacitance oscilloscope probe.

The usual designation for a 10:1 *division* probe is "10X," a practice that may tend to confuse. When a 10X probe is used, the displayed voltage is 1/10 the actual voltage, so multiply the indicated voltage by a factor of 10.

The RC combinations in the circuit can still cause a phase shift, so some manufacturers place a variable trimmer capacitor in the circuit. In low-cost in-

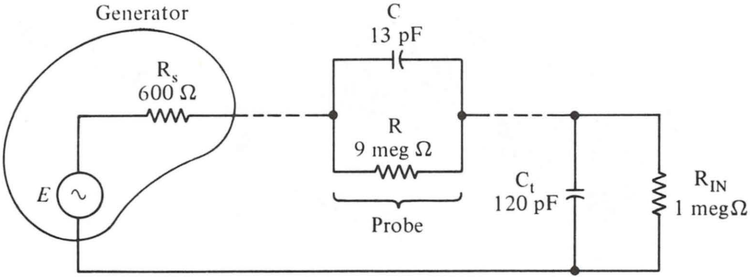


Figure 15-8. Equivalent instrument input circuit when a low-capacitance probe is used.

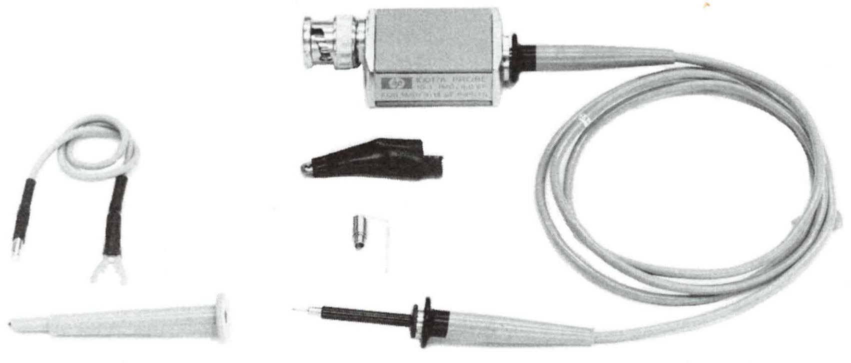


Figure 15-9. Typical low-capacitance oscilloscope probe. (Courtesy of Hewlett-Packard.)

struments the capacitor inside of the probe itself is made variable, whereas in most professional-grade instruments the variable capacitor is located inside the molded BNC connector that attaches to the instrument. This capacitor will eliminate most phase shift problems. Figure 15-10(a) shows the normal waveform,

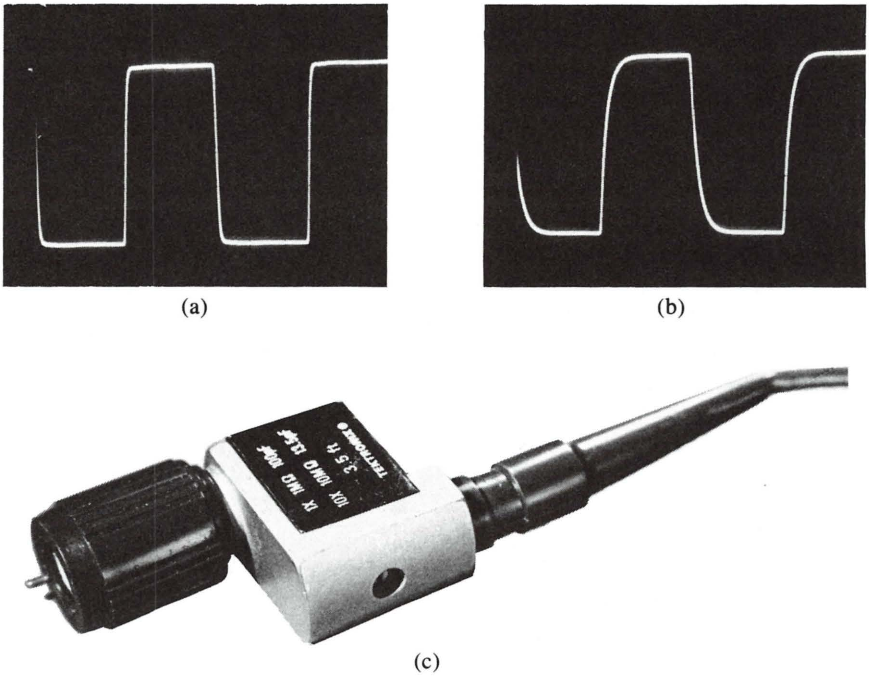


Figure 15-10. (a) Normal trace. (b) "C" misadjusted. (c) Connector end of low-capacitance probe showing access hole to compensating capacitor.

while Fig. 15-10(b) shows the result of a misadjustment—loss of high frequencies. The opposite problem, enhanced high frequencies (i.e., peaking of the waveform), also occurs. Figure 15-10(c) shows the access hole for the adjustment capacitor in a Tektronix probe.

15-9 PROPER PROBE USE

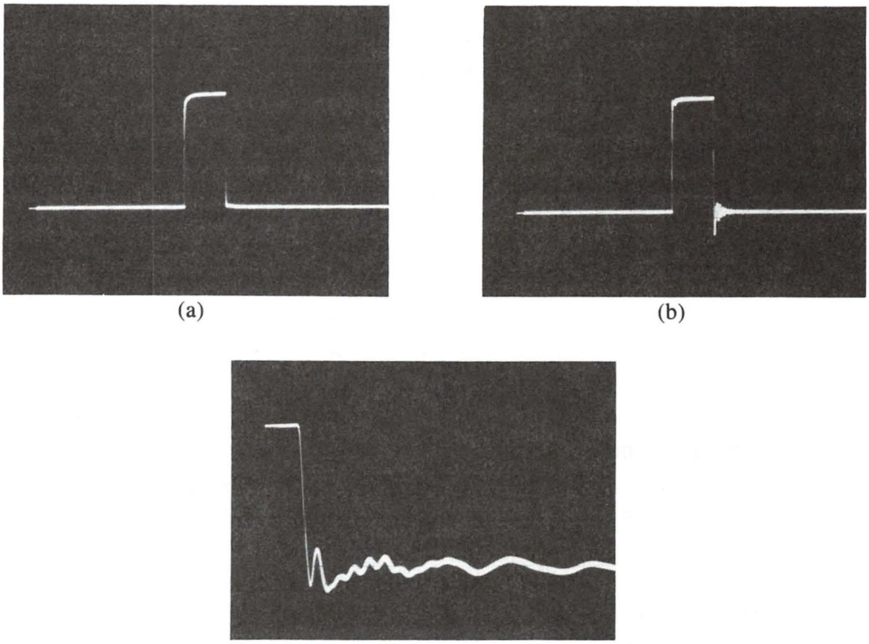
The low-capacitance probe is easy to use, and for the most part produces good results even in inexperienced hands. But there are limitations and pitfalls. Be sure, for example, to heed the manufacturer's frequency response specification, and to set the frequency compensation properly. The fact that a probe was compensated on another instrument is irrelevant.

Some probes have a switch to allow selection of 1X–10X positions. In one position the probe is direct, while in the other it is a divider RC probe. Be sure of which position that switch is set for, or there will be an order of magnitude error in the measurement. This seems like a trivial admonition, but it happens so often that it bears emphasis. In fact, one oscilloscope manufacturer (Tektronix) uses a special BNC connector on its scopes to sense which position the switch is set to, and to indicate the proper volts/div factor either on the CRT screen, or on the vertical attenuator switch.

Another pitfall involves the ground wire on the probe: *use it*. Since most modern instruments use a three-wire power cord from the ac mains, their cabinets are at, or *near,* ground potential. Some people tend to remove ground wires from their probes and then depend upon the ac power mains ground between the instruments to make the connection. Although the ground wire on the probe is sometimes a nuisance, there are at least two faults with this system: *ground loops* and *ringing*.

A ground loop occurs because of voltage drops in the power line ground wiring. The ac currents passing through the ground system cause these voltage drops because of the ohmic resistance of the wires and resistance in connectors. The electrical potential at each chassis, then, is a little different from the potential on nearby chassis. The oscilloscope or voltmeter will see this difference of potential as a valid signal, so an artifact in the measurement is created.

Ringing is shown in Fig. 15-11. The normal pulse is shown in Fig. 15-11(a); the waveform photograph was taken with the probe's ground wire connected properly to the circuit under test. In Fig. 15-11(b) we see the result of *disconnecting* the probe ground, and depending upon the power mains ground to make the return connection. Ringing oscillations follow the leading and trailing edges of the pulse, although it is most pronounced following the trailing edge. This phenomenon is due to the stray capacitances and inductances in the ground circuit, which form a distributed tank circuit. Like any tank circuit energized by a pulse, it will ring, that is to say, oscillate with an exponentially decaying



(a) (b) (c)

Figure 15-11. (a) Normal. (b) Ringing due to lost ground. (c) Ringing portion of (b) expanded.

amplitude. The features of this ringing behavior are shown more clearly in Fig. 15-11(c). This type of artifact is eliminated by the probe ground connection.

15-10 HIGH-VOLTAGE PROBES

Very few oscilloscopes or voltmeters have a maximum dc voltage range in excess of 1500 volts. But many rather ordinary circuits operate at potentials far in excess of this value. A color television receiver, for example, may use up to approximately 25,000 volts on the post deflection accelerator anode.

A high-voltage probe is a special voltage divider (see Fig. 15-12) that allows these high potentials to be read on an ordinary voltmeter or oscilloscope. Most of these probes use the input resistance of the instrument and a 900-megohm resistor inside of the probe as the divider, although a few exist that include the entire divider inside of the probe. For electronic voltmeters with a 10-megohm input resistance, the divider ratio is 100:1, but for most oscilloscopes the ratio is 1000:1 because of the lower (i.e., 1-megohm) input resistance of those instruments.

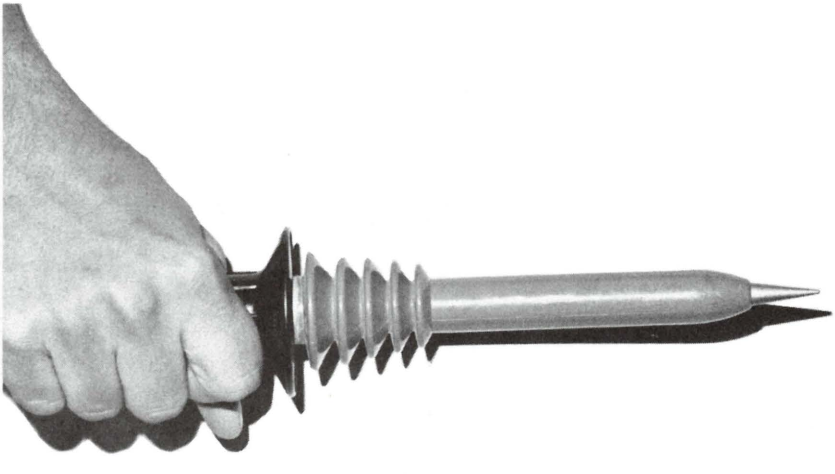


Figure 15-12. High-voltage divider probe.

Note in Fig. 15-12 the construction of the probe; it is built of thick, insulating plastic to prevent electrical shock to the user. The finlike structures on the probe effectively lengthen the *surface path* between the tip of the probe and the user's hand. High tension has the nasty habit of creeping along the surfaces of materials that it cannot go through. High-voltage probes are generally considered safe to use, but must be regarded with caution under some circumstances. For one thing, always use the ground wire attached to the probe, regardless of whether or not the instrument that it is connected to also provides a ground wire. Second, never use a probe that is dirty or wet, or that has grease or other surface contaminants. These could form a surface path for high-voltage arcs to follow. Third, never use a probe that is broken or in poor condition. Additionally, high-voltage probes such as that shown in the figure are not recommended for high-altitude use. At high altitudes, the barometric pressure is lower, so the ionization potential of air is also lower. Equipment used at high altitudes, e.g., mountaintop radar installations or airborne equipment, is specially designed for the application, and this type of probe defeats the special precautions taken by the manufacturer of the equipment.

15-11 RF DEMODULATOR PROBES

Radio frequency voltages cannot be directly measured on most readily available instruments, i.e., multimeters and oscilloscopes. In both cases, a possible solution is the RF demodulator probe shown in Fig. 15-13(a). Diode $D1$ serves to rectify the signal, while the RC network following the diode serves as a filter circuit (as in a power supply). The voltage across capacitor $C2$ will be the peak

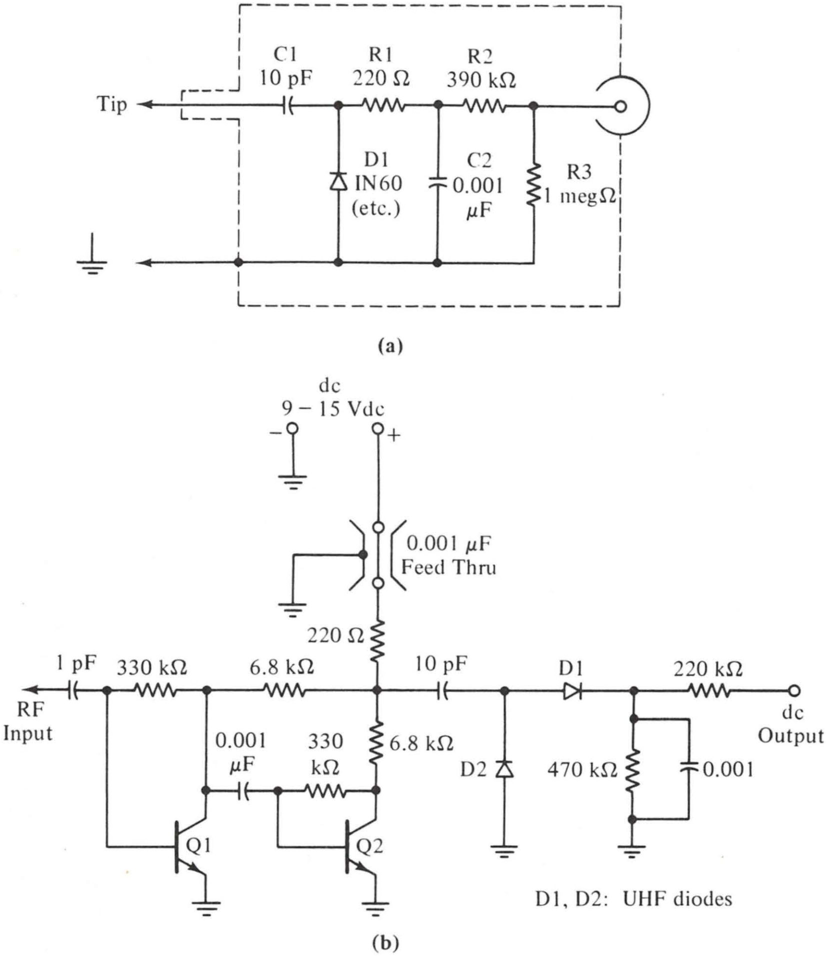


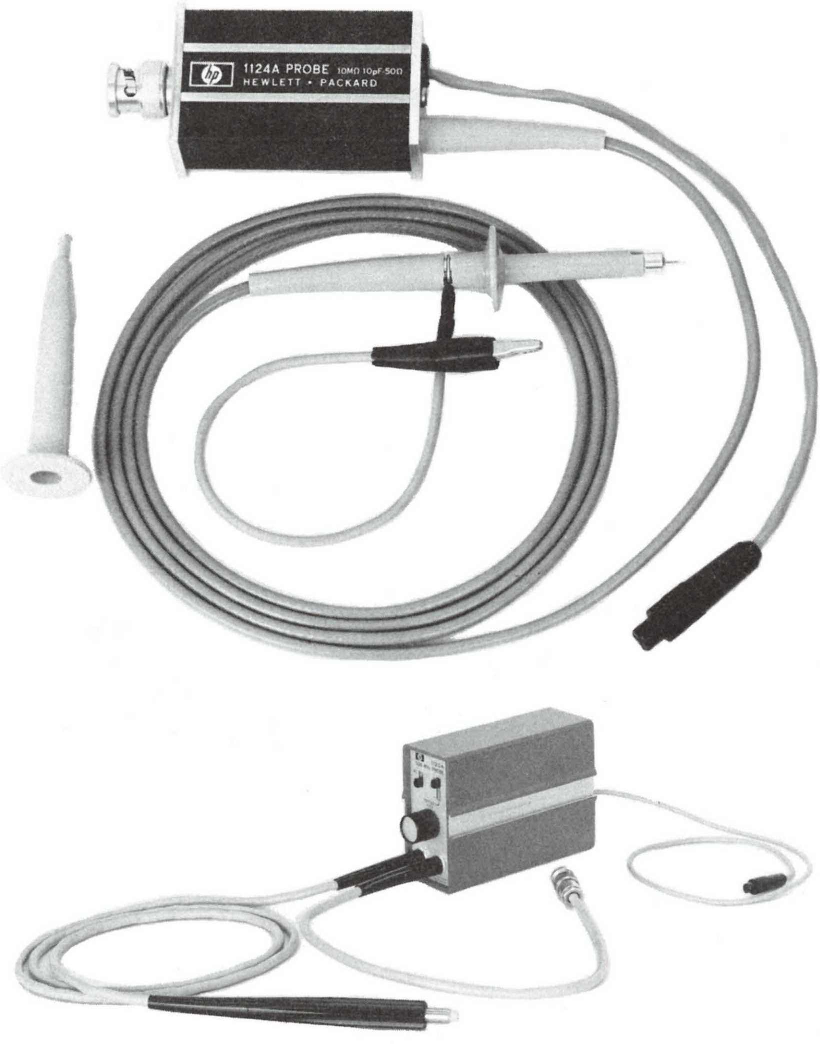
Figure 15-13. (a) RF demodulator probe. (b) Amplified RF demodulator probe.

RF voltage, and the resistor network reduces the peak voltage by 0.707 so that E_0 will represent the rms value, *provided* that the waveform is sinusoidal.

The minimum readable voltage and the maximum peak reverse voltage (prv) is limited by the properties of diode $D1$. If a silicon diode is used, then the 600-mV junction potential is the lowest RF voltage that can be detected, but if a germanium diode is used, then RF potentials down to 200 mV are readable.

The prv rating of most VHF and UHF signal diodes that are suitable for use in an RF probe is between 40 and 100 volts. This can be increased by connecting several identical diodes in parallel, but only at the expense of an increased minimum readable voltage rating.

An *active* RF probe is shown in Fig. 15-13(b). It will amplify weak RF signals before rectification. Transistors $Q1$ and $Q2$ should be selected to have moderate-to-high *beta* ratings, with an F_t of 700 mHz or higher. If proper attention is paid to "good VHF construction practices," then the probe will be essentially flat out to approximately 100 mHz and will operate at reduced gain to approximately 150 mHz. The circuit of Fig. 15-13(b) should be built inside a shielded probe housing. Examples of commercial active probes are shown in Figs. 15-14(a) and 15-14(b).



(a)

(b)

Figure 15-14. (a) Active RF probe. (b) Another active RF probe.

15-12 SPECIAL PROBES FOR IC CIRCUITS

Digital and linear integrated circuits (ICs) are not too forgiving of accidental shorts. If a probe should slip off an IC pin, a very frequent occurrence, then a destructive short may be created. In addition to this problem, the extremely close spacing of IC DIP package pins makes it difficult to attach normal probes. Most often recommended by equipment makers as a solution to this troublesome problem is the clip-on probe shown in Fig. 15-15. This probe is a clothespinlike assembly that has electrode surfaces on the inner sides of the jaws. The operator "squeezes" the grips to open the jaws, and places the electrode surfaces over the IC pins. Multiple clothespin electrodes are often used to troubleshoot or debug digital circuits.

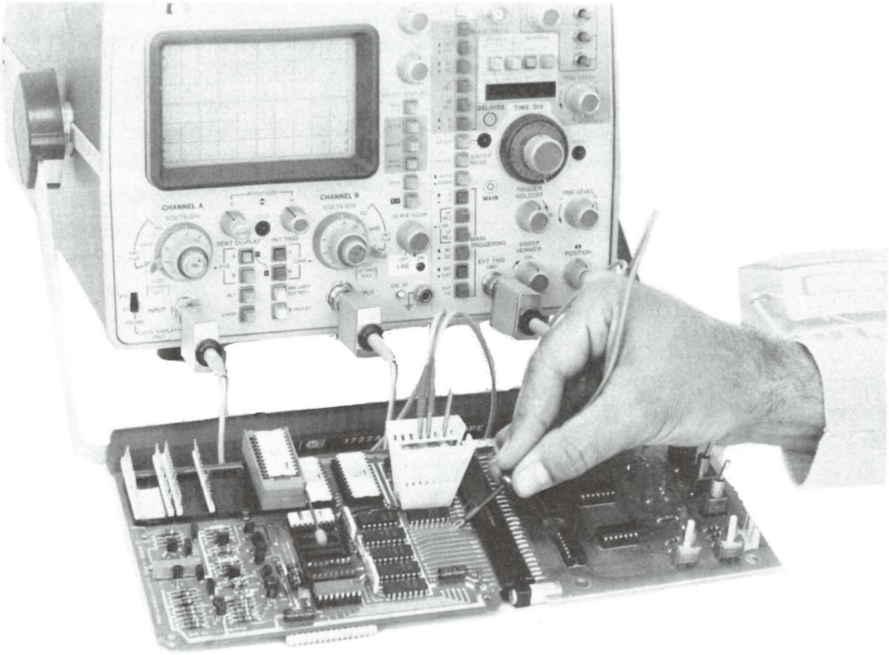
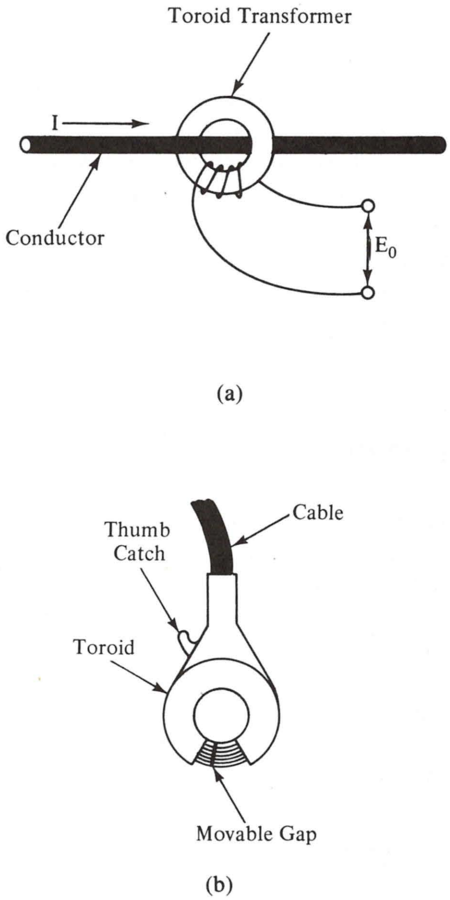


Figure 15-15. Clip-on probe for integrated circuits.

15-13 CURRENT PROBES

The measurement of a *current* by a current meter requires that the circuit be broken so that the meter can be connected in series with the load. But this is not always either possible or desirable. An active probe, such as one of the current-to-voltage converter circuits of Chapter 12 will allow the measurement of dc and low-frequency ac currents on a voltage-measuring instrument such as an oscilloscope or electronic voltmeter. But the requirement for breaking the circuit still exists.

A passive current probe uses the magnetic field surrounding a current-carrying conductor to measure the strength of the current in the conductor. Figure 15-16(a) shows the construction of such a probe, using a toroid, i.e., doughnut-shaped, core that is wrapped with several turns of wire. The conductor carrying an ac current forms the primary of the transformer, while the turns of wire on the



(a)

(b)

Figure 15-16. (a) Current probe circuit. (b) Construction.

core form the secondary. Alternating current in the conductor will induce current in the wire of the probe, so an output voltage E_0 will be generated, and is proportional to the current flowing.

Toroid cores present a problem as far as wrapping around a conductor is concerned. In Fig. 15-16(b) are the constructional details of most typical probes. There is a gap in the toroid. A latch on the side of the core allows the user to spread the gap to admit the conductor. Most of these probes operate over a frequency range of 100 Hz to 3 mHz, but require an output amplifier that produces an output scaled in units of mV/mA.

15-14 SUMMARY

1. Test leads are used at dc and low ac frequencies, where signal amplitudes are relatively high, and circuit impedances are low.
2. Low-capacitance probes are used for general-purpose measurements with oscilloscopes up to approximately 100 mHz.
3. RF demodulator or detector probes are used to make amplitude measurements of RF signals on dc voltmeters.
4. Active current probes use an I-to-E converter circuit, but require series connection in the circuit.
5. Passive current probes can be constructed to simply wrap around the conductor.

15-15 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

15-16 QUESTIONS

1. What is a "probe"?
2. Under what conditions is it allowable to use test leads?
3. Two mechanisms by which test leads pick up interference are ____________ and ____________.
4. A ____________-pair is a type of test lead set that reduces some of the problems associated with normal test leads.
5. List two reasons why the leads in Question 4 are effective.
6. What are the advantages and disadvantages of shielded cable test leads?

7. List two types of connectors used commonly with shielded cable.
8. Which of the two basic connectors is currently favored?
9. What is the purpose of the low-capacitance probe?
10. Most low-capacitance probes have a ____________ division ratio.
11. Measurements may be made with low-capacitance probes using the power mains ground for the return path. True or false?
12. Describe what is meant by a "ground loop" and how it might affect a measurement.
13. A high-voltage probe is actually a well-insulated ____________.
14. Draw a circuit for a demodulator probe that produces an output voltage E_0 equal to the *peak* RF voltage.
15. Describe the differences between a *transmission line* and *shielded cable*. Assume that both are made from RG-58/U coaxial cable.
16. How can VSWR affect the measurement of receiver sensitivity at 150 mHz?
17. What effects may SWR have on frequency counters if the input waveform is a fast rise-time pulse?
18. List two types of active current probe.
19. Most passive ac current probes depend upon the ____________ surrounding the current-carrying conductor.
20. If the load impedance of a transmission line does not approximate Z_0, then a ____________ ____________ device should be used.

15-17 PROBLEMS

1. A signal generator has a 1000-ohm output impedance, and is connected through 10 feet of coaxial cable to an oscilloscope that has a 1 megohm/20 pF input impedance. (a) What is the value of the voltage delivered to the oscilloscope input terminal if the signal source's open terminal voltage is 5.8 V ac? (b) What is the *phase shift* at 10 mHz?
2. The input probe of most VTVMs contains a 1-megohm resistor, while the basic input impedance of the instrument is 10 megohms. Calculate the *percentage of error* created if a technician calibrates the instrument *without* the probe, and fails to realize that the meter markings are valid *with* the probe.
3. Calculate the VSWR of a 27-mHz transmission line if the voltage maxima along the line is 2.8 volts, and the minima is 0.7 volt.
4. A signal generator with a 50-ohm output impedance is connected to a receiver with a 300-ohm input impedance. Calculate the VSWR.
5. A long 50-ohm transmission line is connected to a frequency counter with a 50-ohm input impedance. Calculate the SWR.
6. A 50-ohm source is to be connected to a 100-ohm load through a transmission line. Calculate the surge impedance of coaxial cable that can be

used to make a quarter-wavelength impedance transformer. Make a recommendation as to an appropriate type of cable.

7. How long should the coaxial cable in Problem 6 be if the frequency is 65 mHz and the coaxial has a foam dielectric?

chapter 16

TESTING ELECTRONIC COMPONENTS

16-1 OBJECTIVES

1. To learn how to test pn junction diodes.
2. To learn how to test transistors.
3. To learn how to test capacitors.
4. To learn the types of test equipment on the market.

16-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. What instrument can be used to make a simple qualitative test of a pn junction diode?
2. The same instrument as in Question 1 can also be used to test bipolar transistors. True or false?
3. List three types of transistor tester.
4. Capacitors and inductors can be tested on a ____________ meter.

16-3 TESTING COMPONENTS

It frequently becomes necessary to check individual electronic components either for value (i.e., μF, ohms, beta, etc.) or quality. Service technicians must often use test procedures that determine which component is causing a malfunction. Design engineers often test components to assure critical parameters. The quality control departments of manufacturing plants often inspect incoming components purchased from other manufacturers in order to minimize failures on the assembly line, or product warranty service after their equipment has been sold to the customer.

Some of the tests that we will consider are qualitative in nature, whereas still others are quantitative. Most of the qualitative tests are simple and very easy to perform in a short period of time. Such a test is used mostly in troubleshooting, where we look for gross defects, rather than subtle variations in performance. These tests are valid because of the time they save; why use a curve tracer to determine that a power transistor has a *c-e* short when an ohmmeter will do just as well?

Quantitative tests, on the other hand, are needed to determine actual device performance, or to select out those in a large group that meet tighter specifications than is normally available or guaranteed.

16-4 SIMPLE SEMICONDUCTOR TESTS

A qualitative tester for diodes and bipolar transistors may consist of a simple ohmmeter, while a tester that actually measures the device parameters may tend to be both complex and costly. A solid-state diode consists of a pn junction of semiconductor material. The pn junction will pass current in only one direction, and this forms the basis of a simple test procedure. Figure 16-1 shows the use of an ordinary dc ohmmeter to test diodes. The ohmmeter uses a dc voltage source (usually a dry cell battery) and some calibrating resistors, and measures resistance via the current that is drawn when the unknown resistance is in the circuit. The ohmmeter probes, then, are polarized because the reference voltage is from a dc source; one probe is positive with respect to the other (i.e., "negative") probe. When the ohmmeter is connected as in Fig. 16-1(a), the pn junction of the diode will be forward-biased, so the ohmmeter will register a low resistance; in the range 50 to 1000 ohms in most cases. When the ohmmeter probes are reversed, the pn junction becomes reverse-biased, so the ohmmeter registers a high value of resistance.

The terms "high" and "low" in this context are used in a relative sense. The low resistance is typically in the under-1000-ohms range. The high value tends to be from 5× the low reading in older germanium diodes to greater than 50× the low reading in most modern silicon diodes.

It is important to use the correct ohmmeter scale in this type of test. The use of a too-low scale (that is, ×1 or ×10) can burn out small-signal diodes, while too-high scales tend to give false results. As a general rule, when using an ohmmeter with a 1.5-volt battery, the ×1 and ×10 scales are used for rectifier diodes that are rated at 500 mA or greater forward current. The ×100 and ×1000 scales are suited to testing small-signal diodes (e.g., 1N60, 1N914, etc.).

A problem develops when one is using the ohmmeter sections of the modern digital multimeter (DMM) in this test; the voltage source in the DMM ohmmeter is not a battery, but an electronically regulated power supply. In fact, many manufacturers of these instruments tout these instruments for use in solid-

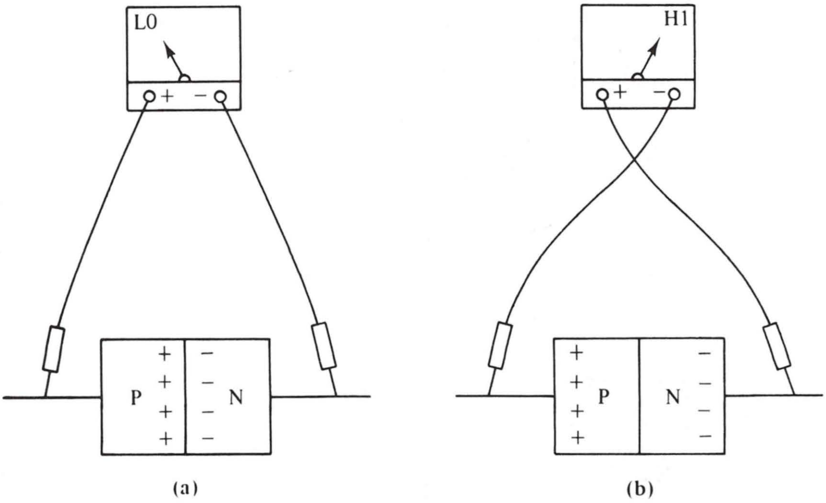


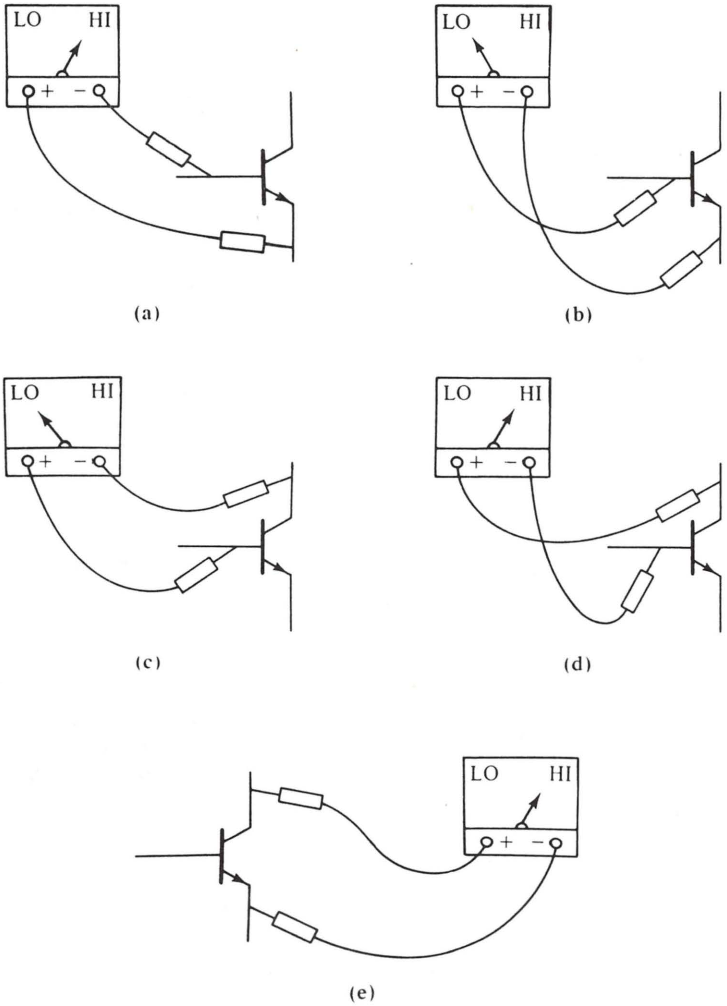
Figure 16-1. (a) Diode shows low resistance when forward-biased by ohmmeter. (b) Diode shows high resistance, or open circuit, when reverse-biased by ohmmeter.

state circuits, because the ohmmeter will not forward-bias pn junctions, or damage anything. This feature is an advantage in that it reduces the chance of false resistance readings in-circuit that are often caused by pn junctions' becoming inadvertently forward-biased. But the low-voltage ohmmeter cannot be used as a tester for pn junctions for the same reason. A few DMM models are equipped with a "high-low power" switch (often labeled "diode" or with the diode circuit symbol) for the ohmmeter specifically so that the user may test pn junctions, while retaining the advantages of low-voltage operation.

A transistor is essentially a pair of pn junctions connected back to back, or at least this is a reasonable model of a transistor for our purposes. We may, therefore, use the same test procedure to test transistors. An ohmmeter, then, may be used to qualitatively evaluate bipolar transistors. Figure 16-2 details the test procedure.

It is necessary to check each pn junction of the transistor separately, and to check the collector-emitter leakage current. Figures 16-2(a) and 16-2(b) show the test to the *b-e* junction, while Figs. 16-2(c) and 16-2(d) show the test applied to the *b-c* junction. Both PNP and NPN transistors are tested in exactly this same manner, but expect the polarity of the high-low readings to be reversed.

Leakage between the collector and emitter terminals is also checked by using the ohmmeter, as shown in Fig. 16-2(e). It is a good practice to measure this resistance by using the reversed probes method, not to discern the quality of a junction, but to keep from forward-biasing one of the other junctions. Use the highest of the two readings as the true reading.



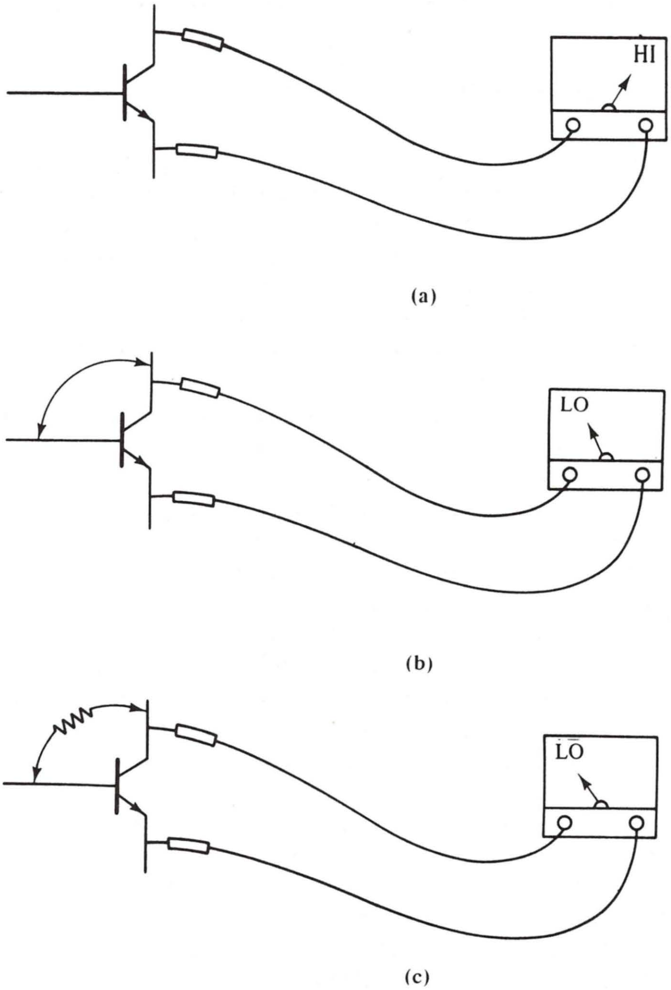
(a) (b) (c) (d) (e)

Figure 16-2. (a) and (b) Checking *b-e* junction of a transistor. (c) and (d) Checking the *c-e* junction of a transistor. (e) Checking *c-e* leakage.

The same advice as to ohmmeter scales must also apply when one is testing transistors. Use the ×1 and ×10 ranges for testing power transistors, and the ×100 or ×1000 scales when testing small-signal transistors.

In testing either diodes or transistors, it is important that you never use an ohmmeter that has a voltage source greater than 1.5 volts. Some older models used batteries such as 4.5 volts, 22.5 volts, or even 45 volts. These voltages *will* burn out the pn junction being tested. Both the voltage and the polarity of the probes can be determined by using a voltmeter to measure the open-terminal voltage that appears across the ohmmeter probes.

The pn junction tests tell us the relative condition of the junctions in the transistor, but do not tell us if the base is able to control the collector current. Those tests are fairly accurate; there are only a few false positives. But a simple test that also uses the ohmmeter, shown in Fig. 16-3, does tell us whether the base terminal is able to control *c-e* current flow. The ohmmeter is connected across the *c-e* terminals in such a way that the *c-e* polarity is proper for normal



(a)

(b)

(c)

Figure 16-3. Checking transistor to see if base can control I_{ce}.

operation. With the base open, there will be a high resistance [Fig. 16-3(a)]. But in Fig. 16-3(b) we see that the *c-e* resistance drops low when the base is shorted to the collector. Some authors, including this one, feel that it is better to connect the base to the collector through a resistor of a value that would normally be used to bias the device (i.e., 30–150 ohms in power transistors, 10–100 kilohms in moderate beta small-signal transistors, and 100 to 1000 kilohms in high beta devices). This will reduce the dependence of the test on interpreting how much of the low resistance reading is due to a forward-biased *b-e* junction.

The ohmmeter tests are generally suited to transistors in the dc to UHF range with the exception of high-voltage types, and VHF power transistors.

16-5 TRANSISTOR TESTERS

There are several varieties of transistor tester on the market: *conduction* (also called *leakage-gain*), *oscillator, beta,* and the *curve tracer*. These different types of tester vary in their relative validity, but all prove useful under the correct circumstances. With the exception of the conduction type, all are considered a step above the simple ohmmeter tests of the previous section. The conduction type of tester is merely a formal version of the ohmmeter tests.

A simple conduction tester, shown in Fig. 16-4, is little more than an ohmmeter and a transistor socket. When the *test* switch ($S1$) is open, the meter will read the leakage resistance between collector and emitter terminals of the transistor. But when $S1$ is closed, we have a rough and purely qualitative measure of the transistor's beta.

The meter in the conduction tester will have two scales; one each for *leakage* and *gain*. Only a few models bother with numbers, which in this case are essentially meaningless, but use color scales red-yellow-green to indicate *bad-who knows-good* (respectively).

The oscillator-type of tester simply places the transistor in an audio or RF oscillator circuit to see if it will oscillate. An ac-coupled detector drives a meter to indicate when oscillations are present. The detector output is zero if the transistor is not oscillating. The oscillator-type of tester not only works with bipolar transistors, but also works (if appropriate circuitry is provided) with both types of field effect transistor.

A true beta transistor tester gives both qualitative and quantitative information about the device under test. Some models are partially automatic, while others, costing very little money, are strictly manual.

The simplest method for measuring the beta of a transistor is to observe that β is the ratio I_c/I_b. In Fig. 16-5 we see one of the simpler circuits used in this type of tester; it is similar to the circuit of the conduction tester. The difference between this and the other type of tester is that the base current is fixed, constant, and known. We may then place a milliammeter in the collector circuit to measure

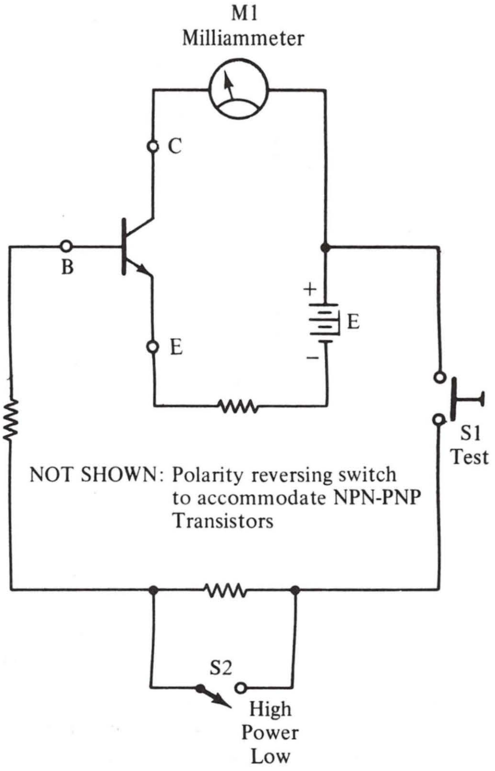

M1
Milliammeter
C
B
E
+
E
−
S1
Test
NOT SHOWN: Polarity reversing switch to accommodate NPN-PNP Transistors
S2
High
Power
Low

Figure 16-4. Conduction-type transistor tester.

the collector current. We are justified in calibrating the collector meter in beta units, because the base current is known, beta is a fixed property of the device, and the collector current is the variable measured. In some models a small ac signal is applied to the base, and the meter is ac-coupled to the collector. This will read ac beta; i.e., h_{fe} instead of H_{FE}.

There are numerous instruments on the market that accurately test transistors for beta. It is generally agreed that a good tester will do the following:

1. Measure I_{cbo}.
2. Measure I_{ceo}.
3. Measure beta in several ranges.
4. Measure super-beta up to 20,000 or more for testing Darlington devices.
5. Test pn junction diodes (both rectifiers and small-signal devices).
6. Test both types of FET.

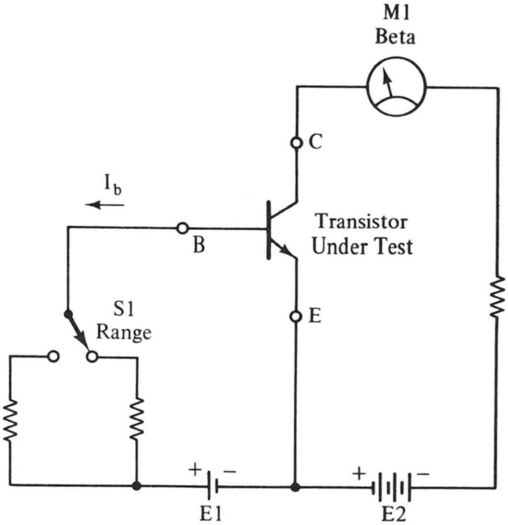
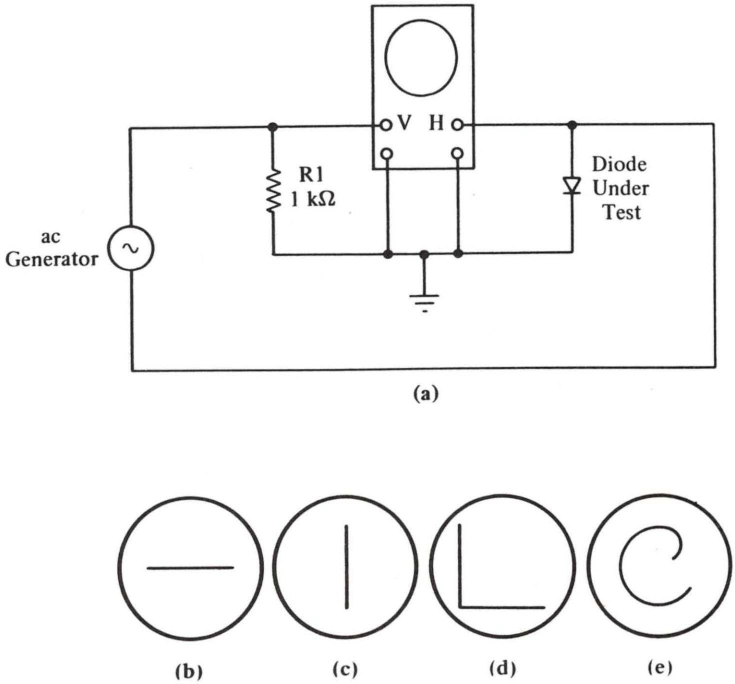


Figure 16-5. Beta tester.

In-circuit transistor testers are capable of approximating the performance of out-of-circuit transistor testers. Most of these are either ac beta testers, though a few are oscillator types. These testers are sometimes guilty of substantial error, so are used for screening of several transistors in the circuit while troubleshooting. A majority of the errors are "false-bad" readings, so one is advised to remove a suspect bad transistor from the circuit for retesting out of circuit. The in-circuit tester appeals mostly to servicers, for whom it saves a lot of time.

The *curve tracer* is an instrument that uses an oscilloscope to trace the characteristic curve of a transistor or diode. These instruments plot the I_c-vs-V_{ce} curve of the device under test at anywhere from one to ten different base current levels. Most of these instruments use step-switched circuits to automatically select the I_b levels, and a variable voltage collector supply to change V_{ce}.

A simple curve tracer for testing diodes is shown in Fig. 16-6(a). In this circuit an ac source is connected between the vertical and horizontal inputs of an oscilloscope. A load is connected across the vertical input, while the diode being tested is connected across the horizontal input of the oscilloscope. If the diode is open, then only horizontal deflection occurs [Fig. 16-6(b)]. A shorted diode allows only vertical deflection [Fig. 16-6(c)]. But a good diode will appear shorted on one-half of the ac cycle, and open on the other half of the ac cycle, because the alternating current alternately forward- and then reverse-biases the diode. This action creates the L-shaped curve of Fig. 16-6(d); it is essentially a combination of the first two traces. A poor diode has a lot of leakage current flowing across the diode even in the reverse-bias direction. Such a diode will produce the trace shown in Fig. 16-6(e).



(a)

(b) (c) (d) (e)

Figure 16-6 (a) Diode curve tracer. (b) Pattern for open diode. (c) Pattern for shorted diode. (d) Pattern for good diode. (e) Pattern for poor diode.

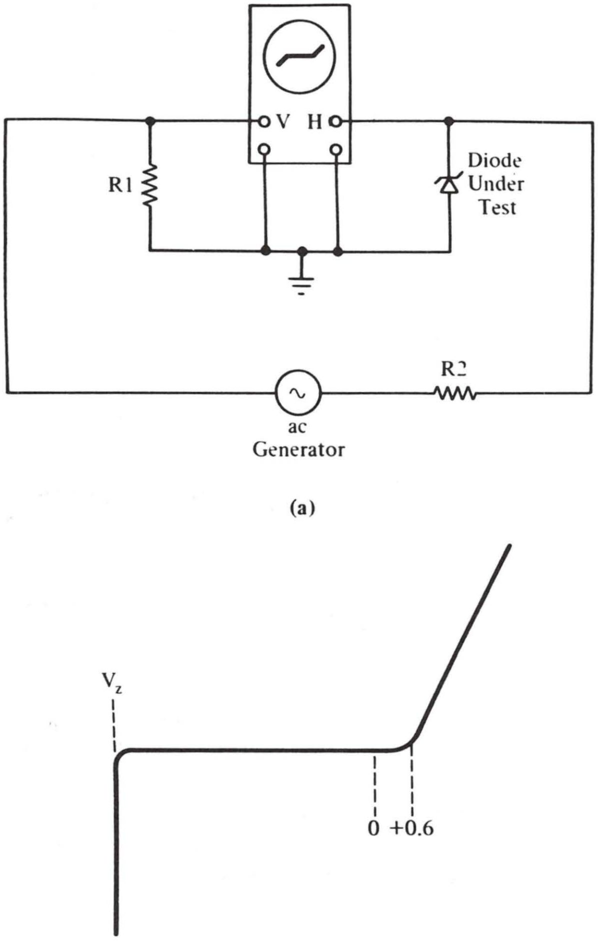
Zener diodes can be tested on a similar curve tracer, as shown in Fig. 16-7(a). The major difference between these two test circuits is the series resistance *R*2. The oscilloscope trace for a good diode is shown in Fig. 16-7(b).

16-6 CHECKING CAPACITORS

Capacitors are simple devices, so (it seems) should be very easy to test. But this is not always true, especially at very low or very high capacitance values. Fortunately, the qualitative tests tend to be very simple.

16-7 TESTING CAPACITORS WITH AN OHMMETER

When you connect a discharged capacitor across an ohmmeter, the capacitor will charge because of the ohmmeter current. The initial current flow is very large, so the ohmmeter will show a low resistance. But as the capacitor charges, the resistance reading will become progressively higher, until it shows an open circuit. At



(a)

(b)

Figure 16-7. (a) Zener diode curve tracer. (b) Curve for good zener diode.

that point, the charge voltage across the capacitor will be equal to the open-terminal voltage of the ohmmeter, so no further current flows. Use the highest ohmmeter range available, or the capacitor will charge too fast for a heavily damped analog meter pointer to deflect properly. This problem becomes especially critical at low capacitance values where the capacitance and the looking-back resistance of the ohmmeter form a short RC time constant. It takes some practice to become adept at judging capacitors of various values by using this technique, so some practice is in order for the student.

16-8 TESTING CAPACITORS WITH A VOLTMETER

We may use a voltmeter in two different ways to evaluate the leakage across the dielectric of a capacitor. Both of these methods are generally used in troubleshooting electronic problems.

In one method, useful in circuits at points where there is a potential of at least several volts across the capacitor, we check for the existence of current flow through the capacitor by looking for a dc voltage at one end. The "cold" end of the capacitor is disconnected from the circuit, and a voltmeter used to measure the voltage from the disconnected end to ground. If there is a problem with leakage, then a voltage will be present.

The second method involves charging the capacitor to some voltage with a power supply. The supply is then disconnected, and a very high impedance voltmeter (i.e., an FET or VTVM type) is used to measure the voltage across the capacitor. A perfect capacitor will hold the voltage indefinitely, but practical capacitors will leak, so the charge diminishes. Again, some practice is needed before an individual can correctly judge whether any given capacitor is "bad" by noting the rate at which the charge decays, so some more practice is in order.

16-9 MEASURING CAPACITANCE

Two traditional methods have been popular for measuring capacitance. One of these is to place the capacitor in one leg of an ac bridge (see Chapter 3). A properly constructed bridge circuit can be calibrated to yield very accurate measurements. The other technique is to use a "Q-meter." This instrument places an unknown reactance (L or C) in a series-tuned tank circuit, and then measures the voltage drop across the unknown element. At any given frequency, the Q-meter can yield an accurate determination of either inductance or capacitance, depending upon how it is configured.

In recent years the bridge and the Q-meter have been somewhat eclipsed, except at the highest accuracy end, by the simple digital capacitance meter. A block diagram of such a meter is shown in Fig. 16-8. The circuit consists of a period counter in which the *on* time of the main gate is set by a monostable multivibrator (i.e., one-shot). The capacitor under test sets the timing of the one-shot stage. In the past, RC one-shot circuits were subject to a lot of error, but today we have IC one-shot circuits that are capable of 1 percent or better accuracy. The period of the one-shot output pulse is equal to the time constant RC. The clock rate in this example is 1 mHz. We find that the count that is accumulated in the counter will be numerically equal to the capacitance in picofarads. For example, if we let R be 1 megohm and a 10-pF capacitor is placed across the test terminals, the gate time will be (10^{-11} F)(10^{6} ohms), or 10^{-5} second. During this time, the 1-mHz clock will input 10 pulses to the counter, so the count is 10. If a 100-pF

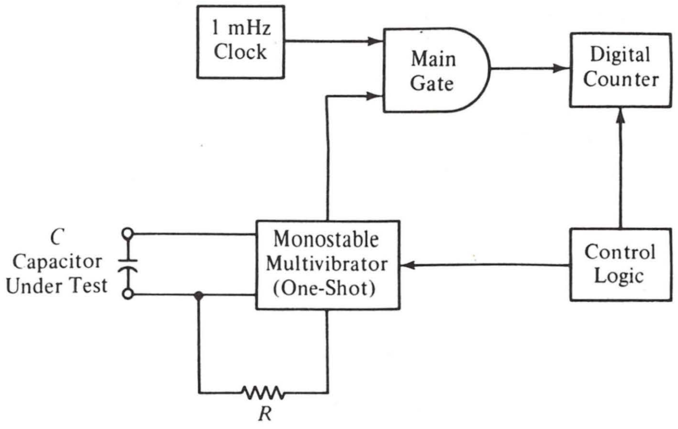


Figure 16-8. Digital capacitance checker.

capacitor were used instead, the gate's *on* time would be 10^{-4} second, so 100 counts of the 1-mHz clock are accumulated. Table 16-1 shows an appropriate timing scheme.

Table 16-1 TIMING VALUES-VS-COUNTS FOR FIGURE 16-8.

$C_{(pF)}$	R *(ohms)*	$T = R \times C$ *(s)*	*Count*
10	1×10^6	1×10^{-5}	10
50	1×10^6	5×10^{-5}	50
100	1×10^6	1×10^{-4}	100
500	1×10^6	5×10^{-4}	500
1000	1×10^6	1×10^{-3}	1000
5000	1×10^6	5×10^{-3}	5000
10,000	1×10^6	1×10^{-2}	10,000

Another approach, although less widely used, is shown in Fig. 16-9. In this case, a voltage comparator circuit is used to gate pulses into an operational amplifier integrator (see Chapter 12). The output voltage of the integrator increments a small amount for each pulse received. One input of the comparator is biased to $0.636E$, while the other is biased by the capacitor charge voltage in the RC timing network. When E_c rises to $0.636E$, i.e., at the end of one RC time constant, the comparator will shut off the flow of pulses to the integrator.

When the integrator input no longer sees pulses, the output voltage will cease to increase. The voltage at this instant is proportional to the capacitance. This technique was popular in predigital instruments because an analog voltmeter, with a scale calibrated in capacitance, could be used as a readout device.

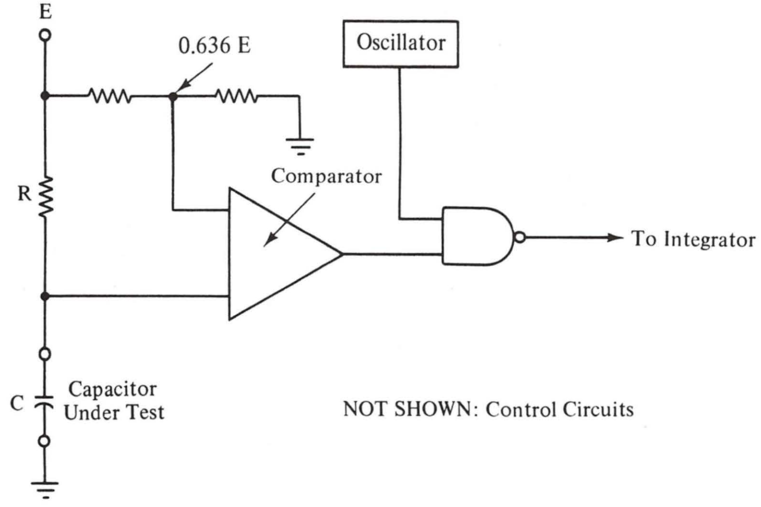


Figure 16-9. Simple analog capacitance tester.

16-10 SUMMARY

1. Qualitative tests of diodes and transistors can be made by using an ohmmeter.
2. Four different types of bipolar transistor tester are *conduction* (i.e., *leakage-gain*), *oscillator, beta,* and *curve tracer.*
3. Capacitors may be checked qualitatively by using an ohmmeter.
4. Capacitance can be measured in an ac bridge circuit, by a Q-meter, or with a digital counter that is controlled by a one-shot in which the capacitor in question sets the timing.

16-11 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

16-12 QUESTIONS

1. An ohmmeter can be used to test a pn junction diode by ____________ the ohmmeter probes and comparing the ____________ and ____________ resistances across the junction.

2. Describe in your own words a procedure for testing bipolar transistors with an ohmmeter.
3. Which ohmmeter scales are appropriate for testing small-signal transistors?
4. Why are some digital multimeter ohmmeter sections *not* suitable for testing pn junctions?
5. List four types of transistor tester.
6. Describe how an ohmmeter can be used to make a qualitative test of a capacitor.
7. Which ohmmeter range should be used for the test discussed in Question 6?
8. List two ways to measure capacitance.

16-13 PROBLEMS

1. A comparator allows pulses from a 1000-kHz clock into a digital counter for one RC time constant. If the resistor is 1 megohm, and the count is 5625, the capacitance is ____________.
2. Find the RC time constant of a 0.001-μF capacitor and a 1-megohm resistor.
3. A bipolar transistor creates a 10-mA collector current when 86 μA flows into the base terminal. Find the *beta*.

chapter 17

MEASUREMENT OF FREQUENCY AND TIME

17-1 OBJECTIVES

1. To learn the different types of equipment used to measure frequency.
2. To learn how to measure time in electronic circuits.
3. To learn frequency measurement techniques.
4. To learn the different types of frequency standards.

17-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. How do frequency and period counters *differ* from each other?
2. What are *Lecher wires?*
3. How can frequency be measured with a WWVB comparator receiver?
4. List three types of absorption frequency meter.
5. List two techniques for using a low-frequency counter to measure a VHF frequency.

17-3 FREQUENCY AND TIME

Frequency is the measure of *events per unit of time*. In the context of electrical circuits the "events" are cycles of alternating current or pulses, and the unit of time is the second, or a subunit such as millisecond (10^{-3} s), microsecond (10^{-6} s), or nanosecond (10^{-9} s).

The unit of frequency is the hertz (Hz), and 1 Hz is equal to 1 cycle per second. In many applications, where the Hz is too small a unit, kilohertz (10^3 Hz) and megahertz (10^6 Hz) are used.

Time that people are used to is measured in hours, minutes, and seconds, and is keyed to the rotation of the earth. In fact, most early astronomical observatories were constructed not for the purpose of conducting scientific research, but to provide the local monarch with the means to keep time. It was the furtherance of commerce and navigation that prompted governments to build observatories; the scientific researches of the astronomer were a benefit.

When dealing with non-dc electrical waveforms we often must consider frequency, period, and duration, all of which are related to time. We have already defined frequency as events per unit of time. Figure 17-1 shows a pulse waveform. The frequency of the pulse train is the number of pulses that occur in one second.

Period and duration are both time measurements, but they are slightly different. Period is defined as the time interval between identical features or points on successive waveforms. The concept of period assumes *periodicity,* that is to say the waveform is not a single event, but repeats itself with predictable regularity. We could, then, specify the period of the waveform in Fig. 17-1 as the interval $(t_2 - t_0)$ or, identically, $(t_3 - t_1)$. The duration, on the other hand, is the time interval during which the pulse is high, i.e., interval $(t_1 - t_0)$ or interval $(t_3 - t_2)$ in Fig. 17-1. Note that duration is not generally considered a feature of sine waves, in which case we consider only period, but is a property of pulse waveforms.

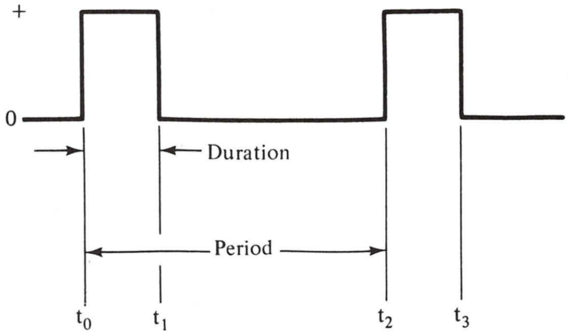


Figure 17-1. Time relationships on pulse.

Frequency and period in repetitive wave trains are related by the expression

$$F = 1/T \tag{17-1}$$

where F = the frequency in hertz (Hz)
T = the period in seconds (s)

Example 17-1

Calculate the frequency of a waveform that has a 225-microsecond period.

SOLUTION

$$F = 1/T \tag{17-1}$$

$$F = \frac{1}{225\ \mu\text{s} \times \dfrac{1\ \text{s}}{10^6\ \mu\text{s}}}$$

$$F = \frac{1}{2.25 \times 10^{-4}\ \text{s}} = \mathbf{4444\ Hz}$$

17-4 PERIOD MEASUREMENT

Period is measured by comparing the event being tested with a known clock or time base. In Chapter 6 we studied the digital counter in which the precision pulses from a time base section were passed through a gate to a decimal counting assembly under command from the event being measured. Counter input circuitry can be adjusted to trigger the gate open and closed, using identical points on successive waveforms.

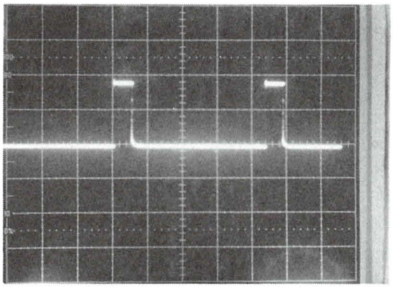


Figure 17-2. Measuring period on oscilloscope screen.

Alternatively, you can often use an oscilloscope time base to graphically measure the period of the waveform displayed on the CRT screen. Figure 17-2 shows two pulses displayed on an oscilloscope screen. The period is 4.4 horizontal divisions, and since the time base was set to 2 ms/div, it represents a time interval of

$$4.4\ \text{div} \times \frac{2\ \text{ms}}{\text{div}} = 8.8\ \text{ms}$$

The limitation on the accuracy of this technique is the accuracy of the oscilloscope time base. If greater precision is required, then an external time base marker generator can be used on the second vertical channel. This generator will place precisely timed marked spikes on the CRT screen, by which time can be measured.

17-5 FREQUENCY MEASUREMENT

Frequency measurements are part of many different electronic and scientific activities. In communications, for example, the Federal Communications Commission (FCC) requires that transmitters be kept on frequency, within a certain error or tolerance band..An AM band broadcast station, for example, must operate on an actual frequency that is ±20 Hz of its assigned frequency. FM band broadcast stations, on the other hand, must stay within ±2000 Hz of their assigned frequency. Class-D Citizens Band transmitters must be within ±0.005 percent (i.e., 50 parts per million—50 ppm). Some two-way radio services must maintain transmitter operating frequency to within ±2.5 ppm.

In most noncommunications electronics, frequency measurement requirements may be greater or less than those in any given communications service. Alternatively, there may be a special measurement problem. An example of a "problem" measurement is sometimes encountered in servicing electronic musical instruments; we may be required to measure a tone to a resolution of two or three decimal points of a Hz, e.g., 680.24 Hz. Many instruments will measure 680 Hz with little difficulty, but will not be able to resolve the fractional cycle, i.e., 0.24, portion. A *high resolution counter* will do the job by actually measuring the *period,* and then taking its reciprocal in calculatorlike circuits before displaying the result on the digital readout. The period of 680.24 Hz, for example, is 1.47×10^{-3} s, or 1.47 ms. A period counter will accurately measure this period, to the required resolution, if it has a 1-mHz time base.

17-6 ROUGH FREQUENCY MEASUREMENTS

Several techniques may be used to make *approximate* frequency measurements, and are useful in those cases where there is no pressing requirement for great precision.

One popular rough measurement is to use the oscilloscope time base to find the period, and then take the reciprocal of period [Eq. (17-1)] to find the frequency. In Fig. 17-2, for example, the period was 8.8 ms, so the frequency is

$$F = 1/T$$
$$F = 1/(8.8 \times 10^{-3}\ \text{s}) = 114\ \text{Hz}$$

Another useful technique (shown in Fig. 17-3) is to connect a calibrated oscillator to an oscilloscope input (the horizontal input is used in Fig. 17-3), and apply the unknown signal to the other input. When the oscillator frequency is adjusted to be an integer (i.e., 1, 2, 3, . . ., n) multiple or submultiple of the unknown frequency, then a stable Lissajous pattern will be seen on the screen of the CRT (see Chapter 8). The frequency of the unknown is found by applying Eq. (8-3).

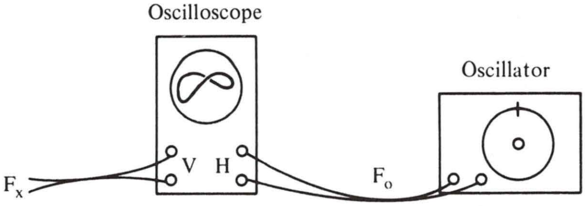


Figure 17-3. Using Lissajous figures to measure frequency.

There are two practical limitations on the use of Lissajous patterns. One, of course, is the precision with which the oscillator is calibrated. In many instances, the oscillator frequency calibration is quite good, while in others it is poor. It is often recommended that the oscillator frequency be measured on a frequency counter, but if a counter is available, then the use of Lissajous patterns is an exercise in the absurd.

The other limitation involves the inability to lock the oscillator on the unknown. The Lissajous pattern will probably *rotate* at a frequency equal to the *difference* between the two frequencies. Trying to make the pattern stand still long enough to make a measurement is a lot like trying to nail Jello to the wall.

It is, however, possible to use the rotation frequency of the trace, and its direction of rotation, to interpolate the correct frequency. The direction of rotation will tell you whether the unknown frequency is above or below the oscillator frequency. This information can be obtained experimentally by rocking the dial of the oscillator back and forth slowly across the correct frequency while observing the Lissajous pattern on the CRT screen. For example, let us say that we have a 1:1 Lissajous pattern that rotates in the direction that indicates that the unknown frequency F_X is *less* than F_0, the known frequency. It is observed that the pattern rotates 26 times per minute. The error is

$$\text{Error} = \frac{26 \text{ cycles}}{\text{min}} \times \frac{1 \text{ min}}{60 \text{ s}}$$

$$\text{Error} = \frac{26 \text{ cycles}}{60 \text{ s}} = \mathbf{0.43\ Hz}$$

Since the unknown frequency is less than the indicated frequency, we subtract the error from the indicated frequency.

$$F_x = 562\text{ Hz} - 0.43\text{ Hz}$$
$$F_x = 561.6\text{ Hz}$$

The usefulness of Lissajous patterns diminishes rapidly as the frequencies involved increase. This problem is due to the severity of certain practical matters, such as the resolution of the RF generator frequency dial.

Another technique for making rough frequency measurements is to use a communications receiver to tune in the unknown signal. We may then obtain a rough indication of this frequency by reading the receiver's dial.

The dial markings of some professional-grade high-frequency (i.e., 3–30 mHz) communications receivers can be quite good. One model, or rather, one series of similar models made since the early nineteen-fifties can claim dial calibration that is accurate to within ±100 Hz. Certain modern solid-state digital models currently on the market not only match the resolution and accuracy of the older models, but will surpass it by an order of magnitude.

Most communications receivers, however, have relatively poor frequency dial calibrations. Even these receivers provide a rough indication, and they can be made substantially more accurate if the dial markings can be compared with an external frequency standard, or one of the *National Bureau of Standards* (NBS) standard frequency-time radio broadcast stations. These stations, using the call letters WWV and WWVH operate on frequencies of 2.5, 5.0, 10.0, 15.0, 20.0, and 25.0 (WWV only) mHz. Alternatively, the dial markings may be compared with a 100-kHz or 1000-kHz crystal oscillator that was calibrated against WWV/WWVH.

The crystal oscillator method can be made very accurate, as will be demonstrated in a later section of this chapter, but only when the receiver dial markings are used only as a guide. Another source must be the standard of measurement.

17-7 ABSORPTION WAVEMETERS

Figure 17-4 shows two elementary forms of absorption wavemeters. These instruments provide only the roughest indication of frequency, but are useful for some purposes, such as determining whether or not the signal a transmitter is producing is a harmonic, parasitic, or the fundamental.

Both types of absorption frequency meter consist of a resonant tank circuit, i.e., a fixed inductance and a variable capacitor. The capacitor is equipped with a dial that is calibrated in units of frequency. The coil in both cases is mounted in a manner that allows it to be coupled closely with the circuit or component being tested. The capacitor, however, must be mounted inside of a shielded enclosure

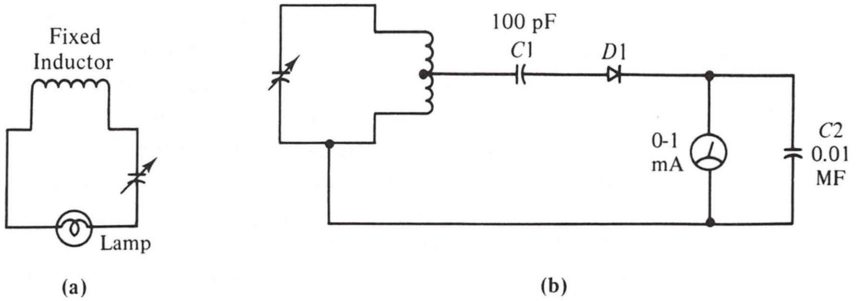


Figure 17-4. (a) Simple absorption wavemeter. (b) Rectified absorption wavemeter.

so that it will not be detuned by stray capacitances (including the operator's own hand). As a result, it is common to find the capacitor mounted inside an aluminum shield box. The inductor, then, is mounted on an insulated socket on the shield box.

The circuit in Fig. 17-4(a) is the classic absorption wavemeter. The capacitor, the inductor, and a lamp form a series-resonant tank circuit, so it will have a very low impedance at the resonant frequency. Energy from the tank circuit of the radio transmitter being tested is induced into the wavemeter circuit through the coil, which is placed in close proximity to the transmitter's tank. This will set up an oscillating current in the wavemeter tank circuit that causes the lamp to light up. It takes a relatively large amount of current to light the lamp, so it is prudent to use low-current types in this application.

In some cases, such as vacuum tube transmitters that have a plate or cathode milliammeter, then the lamp may be omitted. If the inductor of the wavemeter is coupled to the inductor in the transmitter's *plate* tank, then some of the plate energy will be absorbed by the wavemeter. When the two tank circuits are tuned to exactly the same frequency, then the amount of energy absorption increases dramatically. If the dial of the wavemeter is tuned slowly back and forth across the transmitter tank's resonant frequency, then there will be a momentary inflection of the milliammeter caused by the dramatic change in energy absorption level.

A more sensitive variation on the wavemeter circuit is shown in Fig. 17-4(b). In this case, the tank of the wavemeter is used in the parallel resonant mode. Signal from an impedance-matching tap on the inductor is coupled through capacitor *C*1 to the diode, where it is rectified. The pulsating dc waveform is filtered and smoothed by capacitor *C*2. This produces the steady dc current required by the meter.

The milliammeter in Fig. 17-4(b) is deflected an amount that is roughly proportional to the strength of the applied signal. This wavemeter, then, will also function as a *relative* field strength indicator to aid in tuning the transmitter.

The use of an impedance-matching tap on the inductor is necessary because of the low impedance of the diode rectifier. Ordinarily, tank circuits should have a high impedance so that the Q is increased. Also, recall that maximum power transfer between two circuits, or between portions of a single circuit, occurs only when the source and load impedances are equal. Furthermore, the low impedance of the diode would load the tank circuit too much, thereby lowering the Q. The effect of lowered Q is to broaden the tuning, thereby obscuring the resonant point. In some models, there is a link-coupled secondary coil going to the diode circuit, but in this case the main tuning inductor is used as a tapped auto-transformer.

The absorption wavemeter is a very crude device, no matter how well it is constructed. It is also limited in sensitivity. A form of active wavemeter is the *dip oscillator*. If an L-C tank circuit is used as the frequency-determining network in a variable oscillator, then a dip oscillator type of wavemeter is made. Older models used vacuum tubes as the oscillating active element, so were called "grid dip oscillators." Modern dip instruments are called either *base* or *gate dip oscillators,* depending upon whether a bipolar or JFET transistor is used as the active element. One model, using a tunnel diode, was sold under the name of "tunnel dipper."

A typical grid dip oscillator is shown in Fig. 17-5. This circuit is a variable-frequency oscillator (VFO) of the Colpitts type. The inductor from the resonant tank circuit is mounted external to the cabinet housing the rest of the circuit. This is done so that the coil may be easily coupled to the tank circuit or component being tested. The grid current of the oscillator tube is monitored by microammeter $M1$.

Any dip oscillator works because energy from its tank circuit is absorbed by any closely coupled conductor. If that conductor is a resonant tank circuit, however, the amount of energy absorbed will be greatest when the oscillator frequency and the resonant frequency of the tank are the same.

Capacitor $C1$ is coupled to a dial calibrated in units of frequency. The operator holds the inductor on the dip meter close to the circuit being tested, and *slowly* tunes $C1$ until a pronounced dip in grid current is noted. The resonant frequency of the tank circuit is then read from the dial of the dip oscillator.

The chief advantage of the dip oscillator over the simple absorption wavemeter is that it is active, so is, therefore, more sensitive (i.e., it will detect even weak signals). The dip oscillator may also be used on turned-off circuits in which no power is present, because the active element provides the power needed for the test.

The dip oscillator may also be used with energized circuits in a manner that is very similar to the techniques using absorption wavemeters. In that case, we are using the dip oscillator as an *oscillating detector*. A pair of earphones is plugged into jack $J1$, and it takes the place of the meter movement. The coil is coupled to the circuit under test, and $C1$ is adjusted. When the dipper frequency

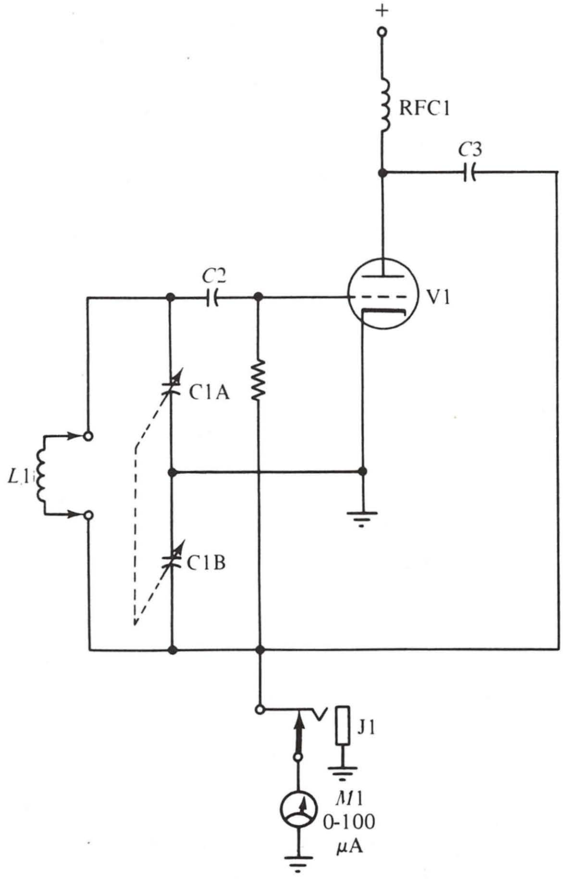


Figure 17-5. Grid dip oscillator.

gets close to the frequency of the signal in the tank circuit being tested, an audio tone is heard in the headphones. This tone is a heterodyne between the oscillator's frequency and the unknown frequency in the tank circuit. Whenever two frequencies mix together in a nonlinear circuit, there are extra frequencies generated equal to their sum and difference. In other words, if two frequencies are mixed together, then there will be *at least* four frequencies produced. If we call one frequency A and the other B, then the four output frequencies will be A, B, $A + B$, and $A - B$. Of these, the first three will be higher than audio, but the difference frequency will be an audio tone. The exact pitch of the tone, then, is equal to the difference between the two frequencies, so when the two frequencies are equal (i.e., $A = B$), then the difference frequency $(A - B)$ will be zero. The two frequencies are said to be in "zero-beat" when they are equal.

None of the absorption techniques is a precise method for measuring frequency. They are used only in cases where simplicity and low cost count for more than accuracy. In transmitter work, for example, we might want to know whether an output signal is due to fundamental frequency oscillation of an amplifier, harmonic generation (i.e., $2F$, $3F$, etc.), or a parasitic that is not harmonically related to the fundamental frequency F.

17-8 LECHER WIRES

Lecher wires are a mechanical device that is used for determining the *wavelength* of signals in the range of frequencies between approximately 50 and 500 mHz. Above that limit, similar techniques involving slotted lines (Sec. 17-9) are used. The slotted line works at frequencies well over 1000 mHz.

A set of Lecher wires consists of a pair of large-diameter parallel conductors, such as copper tubing, or copper wire of #12 size, or larger. The length of the conductors must be at least one-half wavelength at the lowest frequency to be tested. This partially explains the lack of popularity for Lecher wires below 50 mHz, because lengths become too great.

Parallel to the conductor pair will be a measuring scale calibrated in units of length, usually meters, centimeters, and millimeters. A few units might be found that use English units of length.

Some Lecher wires have an RF voltmeter, i.e., a diode detector and a microammeter, such as that shown in Fig. 17-6. This is not strictly necessary, however, if the RF source being tested is a vacuum tube model equipped with a dc milliammeter in the plate or cathode circuits.

A shorting bar between the two conductors is designed so that it can be moved along the length of the conductor from one end to the other. The effective electrical length of the Lecher wires is the distance between the input end and the shorting bar. When one is performing a frequency measurement, the shorting bar is placed initially at the 0-cm mark, and is then moved slowly down the length of the parallel conductors while the RF voltmeter is being monitored. When the bar is moved to the point representing one-half wavelength of the applied frequency, then a pronounced dip is noted in the RF voltmeter reading. The distance between this point and the 0-cm point is equal to the half-wavelength of the applied signal. Some authorities claim that better accuracy results from noting this distance, and then continuing the short further down the wires until another dip is noted. Make an average of the two distances, i.e., 0 cm to first dip and first dip to second dip, and then use the resultant average distance in the standard formula:

$$F = \frac{C}{2\lambda} \tag{17-2}$$

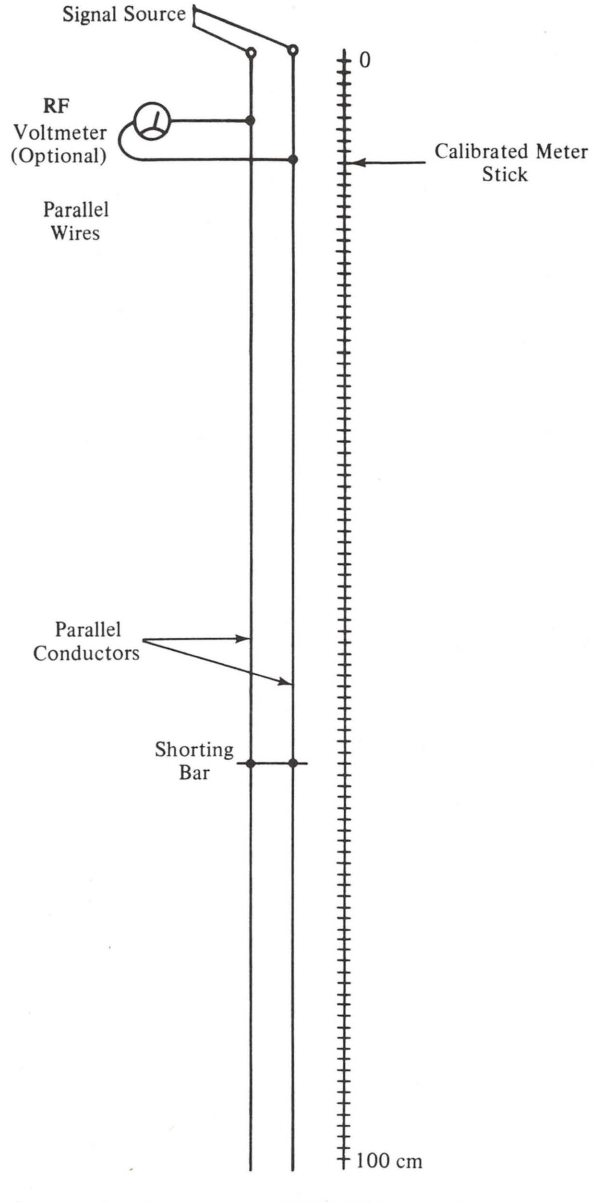


Figure 17-6. Lecher wires for measuring VHF/UHF frequencies.

where C = the speed of light (3×10^8 m/s)
F = the frequency in hertz (Hz)
λ = the wavelength in meters (m)

This equation is also written as

$$F_{(\mathrm{Hz})} = \frac{3 \times 10^8}{2\lambda_{(\mathrm{m})}} \tag{17-3}$$

Since most frequency measurements using Lecher wires are in the over-50-mHz range, the units hertz are too cumbersome, especially given the lack of precision inherent in the Lecher wire technique. We may rewrite Eq. (17-3) to use megahertz instead of hertz, as follows:

$$F_{(\mathrm{mHz})} = \frac{300}{2\lambda_{(\mathrm{m})}} = \frac{150}{\lambda_{(\mathrm{m})}} \tag{17-4}$$

and if English units are used, i.e., *inches:*

$$F_{(\mathrm{mHz})} = \frac{11{,}811}{2\lambda_{(\mathrm{in})}} = \frac{5906}{\lambda_{(\mathrm{in})}} \tag{17-5}$$

In either case, however, we may find an expression for using the full distance between 0 cm and the second dip by eliminating the denominator 2 in Eqs. (17-4) and (17-5).

Accuracies on the order of 1000 ppm are possible, if the operator of the Lecher wire set takes great care, and makes several measurements that are averaged together. Great care in observing the nulls, care in recording the distance data, loose coupling to the source (tight coupling broadens the null), and averaging several measurements are the key to accuracy in Lecher wire use.

17-9 SLOTTED LINE MEASUREMENTS

A technique similar to the Lecher wires is the slotted line detector of Fig. 17-7. The slotted line is a special section of coaxial transmission line that has a slot in its outer conductor running most of its length. An RF detector probe is fitted through the slot, and is coupled to the center conductor by mutual inductance. The output of the detector is a dc voltage proportional to the intensity of the signal in the line at that point.

Some slotted line assemblies have a knob and worm-gear mechanism that allows precise adjustment of the detector position to any position along the slot. A metric scale calibrated in millimeters tells the operator how far along the line the probe is located. A micrometer dial is usually provided, because it is neces-

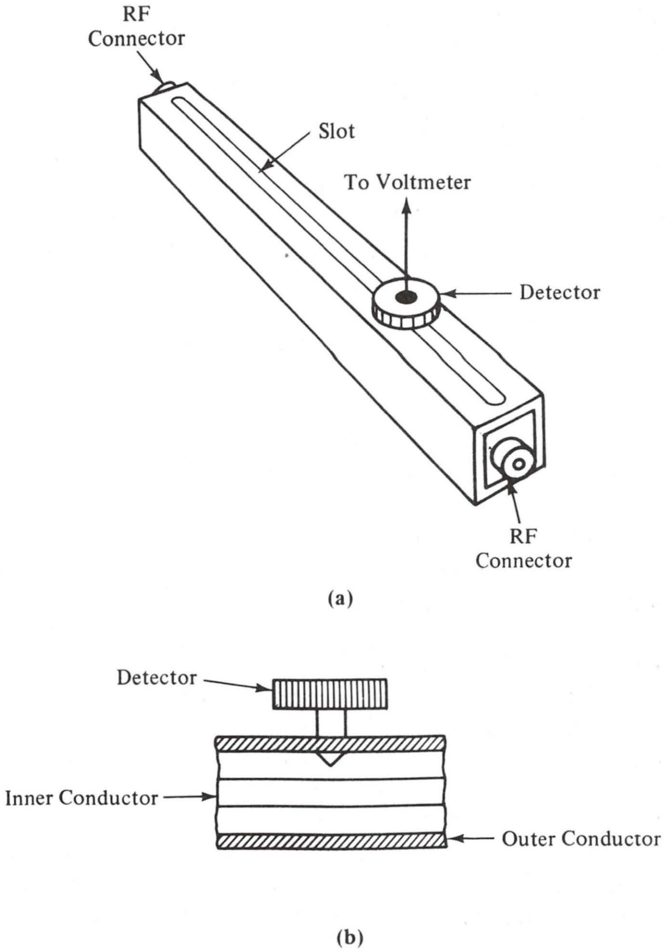


Figure 17-7. (a) Slotted line. (b) Close-up of detector probe.

sary to have a more precise resolution of distances at the higher frequencies covered by the slotted line, compared with the Lecher wire set.

A slotted line is used in a manner that is nearly identical to that of the Lecher wires. Reasonably accurate wavelength (hence frequency) determinations are possible at any frequency above the frequency at which the slot is one-half wavelength. The detector will locate nodes and antinodes along the length of the line, and it is known that nodes are one-half wavelength apart, as are antinodes. The nodes, incidentally, are usually more sharply defined, so are best used to make the measurement. The formulas given earlier for Lecher wires also apply to the slotted line.

17-10 NBS RADIO BROADCAST SERVICES

The *National Bureau of Standards* operates standard frequency and time broadcasting stations. One of these stations, WWV, operates from Fort Collins, Colorado (near Boulder), and the other, WWVH, from near Kekaha on the island of Kauai (Hawaii). WWV broadcasts on frequencies of 2.5, 5.0, 10.0, 15.0, 20.0 and 25.0 mHz, while WWVH operates on all of these frequencies except 25.0 mHz. The operating frequencies of these stations are maintained at ±2 parts in 10^{11} by an atomic clock frequency standard.

The two stations also broadcast audio reference tones. A 440-Hz tone (i.e., musical A-above-middle-C) is broadcast one minute past every hour by WWVH and two minutes past every hour by WWV.

WWV broadcasts a 600-Hz tone during *odd* minutes of the hour, while WWVH broadcasts the 600-Hz tone during the *even* minutes of the hour. In both cases, a 500-Hz tone is broadcast during the alternate minutes. Each audio tone is of 45 seconds duration.

Voice time announcements in Universal Time (UT) are made every minute. WWV uses a male announcer, while WWVH uses a female, so that the listener can recognize which station is being received.

A time marker "tick" is transmitted continuously, except for the twenty-ninth and fifty-ninth seconds of each minute.

Geoalerts are broadcast during the nineteenth and forty-sixth minutes of each hour, and are updated at 0400 UT every day. Voice radio propagation predictions are broadcast during the fifteenth minute of every hour, giving expected high-frequency propagation over the North Atlantic path, considered to be one of the most disturbance-prone radio paths in the world.

Radio station WWVB, also in Fort Collins, Colorado, operates on a frequency of 60.0 kHz. This station also broadcasts a serially encoded binary signal giving the time in UT. The binary time signal is used to automatically update clocks.*

17-11 USING NBS BROADCASTS

The NBS radio broadcasts can be used to compare local frequency standards. The high-frequency broadcasts (that is, 5–25 mHz) can be used to calibrate crystal oscillator standards and some frequency counter time base oscillators to a moderate degree of precision. These oscillators usually operate on a frequency of 100 kHz or 1000 kHz, so harmonics will coincide with a WWV/WWVH frequency. When the two signals, i.e., NBS and crystal oscillator, are both present in the

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*Additional information on WWV, WWVH, and WWVB is contained in *NBS Publication No. 236: NBS Frequency & Time Broadcast Services* (25¢ from the Superintendent of Documents, Washington, D.C., 20402).

receiver output stages, then a heterodyne beat note is created, and it has a frequency equal to the difference between the two input frequencies. The frequency trimmer on the crystal oscillator (usually a variable capacitor in series with or parallel across the crystal element) is then adjusted until the two frequencies are zero-beat.

But the meaning of "zero-beat" varies with different types of measurement apparatus and from one observer to another. A normal human adult, with normal hearing acuity, can hear tones down to 25–30 Hz. This fact, coupled with lower-end bandwidth limitations on the receiver audio/detector circuits, limits the accuracy of the aural zero-beat to ±30 Hz, and may be ±100 Hz in some people.

The receiver S-meter can sometimes be used as a zero-beat detector. It is noted that the S-meter may vibrate back and forth at the beat-note frequency. As the two input frequencies become closer together, the beat note drops in frequency, and diminishes to zero when the two frequencies are identical. In actual fact, it is practically impossible to attain a 0-Hz beat note. But beat notes of less than 1 Hz have been observed by the author, using the S-meter "bobble" technique of observation.

Accuracy of this method is improved by using the *highest* WWV frequency that produces a *strong* signal in your locality. Weak signals are not as useful, because certain other errors become more pronounced. The differences in the results obtained using weak and strong signals from WWV/WWVH suggest that it is prudent to wait until one of the highest frequencies becomes loud in your area—a function of radio propagation conditions.

Let us assume, for purposes of an example, that a receiver with an S-meter or external phase detector is used to zero-beat a 500-kHz time base oscillator to WWV, and that a ±2-Hz accuracy in the adjustment is obtained. If we zero-beat against 2.5 mHz WWV, then we are using the fifth harmonic of the 500-kHz oscillator signal to create the beat note. The ±2-Hz error, then, is now *divided* by the same factor as the harmonic used, so the nearest that we can actually set the oscillator frequency is 500 kHz ±(2/5) Hz, or 500 kHz ±0.4 Hz. This means that the actual frequency of the oscillator will be between 499,999.6 Hz and 500,000.4 Hz. As striking as this accuracy appears, however, compare it with the results obtainable if the 25-mHz WWV frequency is used instead of 2.5 mHz. In that case, we still can zero-beat to ±2 Hz, but the error is now ±2/50 because we are using the fiftieth harmonic of the oscillator to beat with WWV. The error in this case, then, is ±0.04 Hz, a full order of magnitude better.

WWVB offers what is, perhaps, the best frequency comparison, even though it operates at a frequency of only 60 kHz. This statement stands in seeming contradiction to the advice given earlier, but is nonetheless true. The VLF station WWVB produces a signal that is less susceptible to fading, ionospheric disturbances, and the phase shift errors that affect the HF signals. The propagation phase shifts of WWVB are more constant and predictable in any given location than are those for WWV and WWVH.

Figure 17-8 shows the block diagram of a 60-kHz WWVB *comparator receiver,* several brands of which are marketed. The antenna is a shielded loop tuned to 60 kHz, and in most cases includes a tuned 60-kHz amplifier at the feedpoint of the loop. A high-gain 60- to 120-dB amplifier, also tuned to 60 kHz, usually follows the loop antenna. The output of this amplifier chain is usually made available to the user through a front panel connector.

Following the high-gain amplifier is a level detector, or similar stage, to produce a square-wave output for each cycle of the 60-kHz signal. This signal will then be used to drive a phase detector. (Older models did not use the squarer stage, and applied the signal to an L-C/diode phase detector. Modern models use digital coincidence-type phase detectors.)

The frequency being measured is passed through a divide-by-N frequency counter before being applied to the other input port of the phase detector. In some cases, a divide-by-N counter is also used between the output of the squarer and the 60-kHz input of the phase detector.

The output of the phase detector coincidence gate is a series of pulses of widths proportional to the difference in frequency or phase of the two signals at the input ports, so when these pulses are integrated, it will produce a dc voltage proportional to the difference. This voltage is measured on a zero-center galvanometer or a strip-chart recorder.

The unknown frequency can be measured by adjusting the divide-by-N counters until a zero difference frequency indication is obtained. The frequency of the unknown frequency is given by the expression

$$\frac{6 \times 10^4 \text{ Hz}}{N_1} = \frac{F_X}{N_2} \tag{17-6}$$

which can be algebraically rearranged to solve for F_X:

$$\frac{N_2}{N_1} \times (6 \times 10^4) \text{ Hz} = F_X \tag{17-7}$$

where F_X = the frequency of the unknown signal, in hertz (Hz)
N_1 = the division ratio applied to the 60-kHz signal
N_2 = the division ratio applied to the unknown signal

The use of the comparator receiver is a little clumsy for ordinary frequency measurements, but becomes extremely attractive for measuring the long-term stability of counter time-base oscillators.

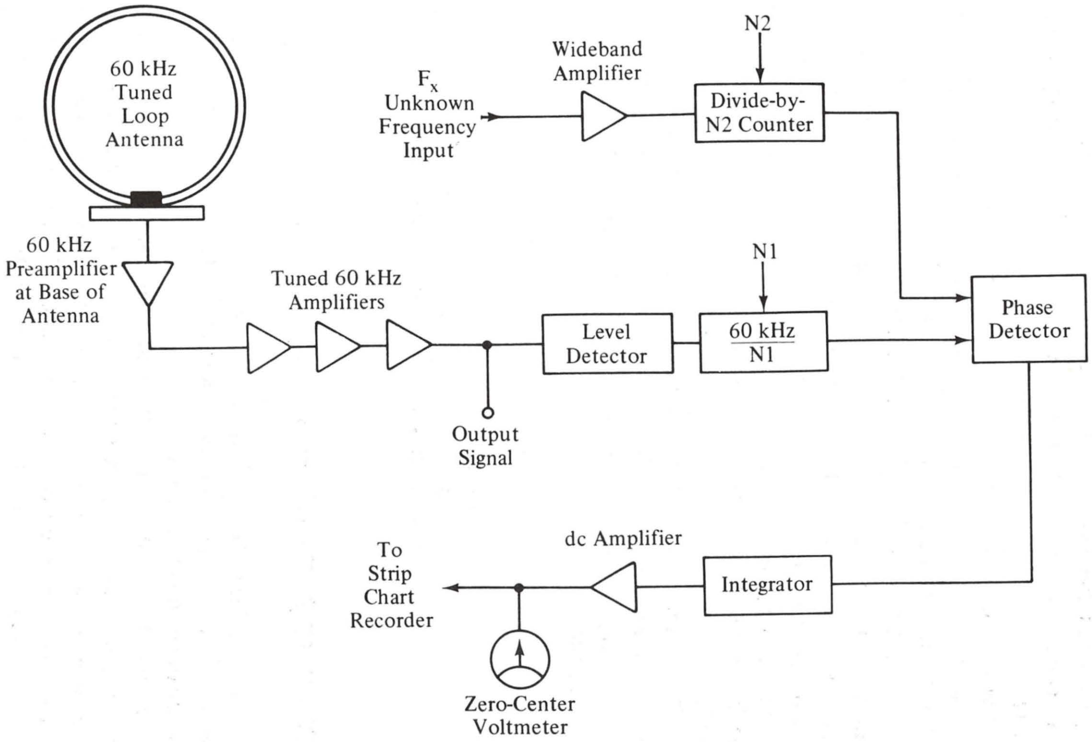


Figure 17-8. Block diagram of a 60-kHz comparator receiver.

17-12 ATOMIC FREQUENCY AND TIME STANDARDS

Until recent decades the earth's rotation was used as the reference standard for the measurement of frequency and time. Frequency was sometimes measured by having the signal run a synchronous electric clock, so by observing how the clock lost or gained time we would know whether or not the frequency was precisely accurate.

But it turns out that the earth's rotation is not the constant it was once believed to be. In fact, there are even seasonal variations thought to correlate highly with the waxing and waning of the polar ice caps.

In 1964, the Twelfth General Conference of Weights and Measures redefined time and frequency in terms of the *atomic clock,* which contains a phase-locked loop (PLL) in which a reference frequency source is either a rubidium gas cell device, or an atomic beam of thallium or (more frequently) cesium atoms. These clocks lock the frequency of a 5- or 10-mHz crystal oscillator to the oscillation frequency of the atoms. It has been found that accuracies of $\pm 7 \times 10^{-12}$ are possible.

The atomic resonator works on the natural resonant frequency of the atom used, which when excited will oscillate (i.e., wobble) at an extremely stable fixed frequency. Table 17-1 shows the oscillation frequency of some atoms often used in atomic clocks.

Table 17-1

Hydrogen maser	1420.405751 mHz
Cesium beam	9192.631770 mHz
Rubidium gas cell	6834.682608 mHz

17-13 BASIC FREQUENCY METERS

There are two basic types of frequency meter in common use: *heterodyne* and *counter*. Although each class has numerous different examples, there are some commonalities among them all.

Figure 17-9 shows the elementary heterodyne frequency meter. In all heterodyne techniques, the unknown signal is mixed with a known signal from a precisely calibrated frequency source. The frequency of the beat note gives the frequency difference between the two sources. If the source is zero-beat with the unknown signal, then the unknown will be equal to either the known frequency or one of its harmonics or subharmonics.

Frequency counters, examples of which are shown in Figs. 17-10(a) and 17-10(b), were discussed in detail in Chapter 6. The reader may find it prudent to

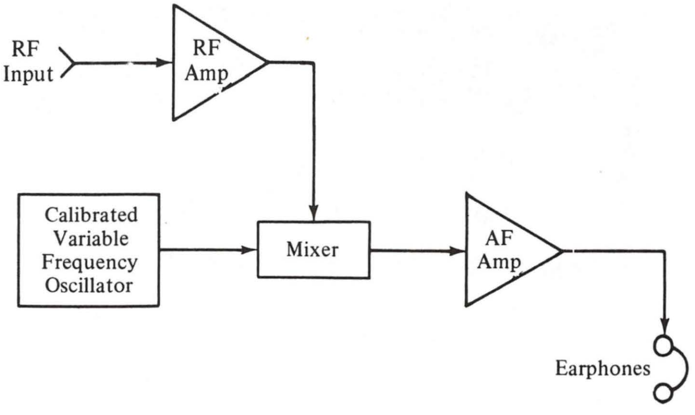
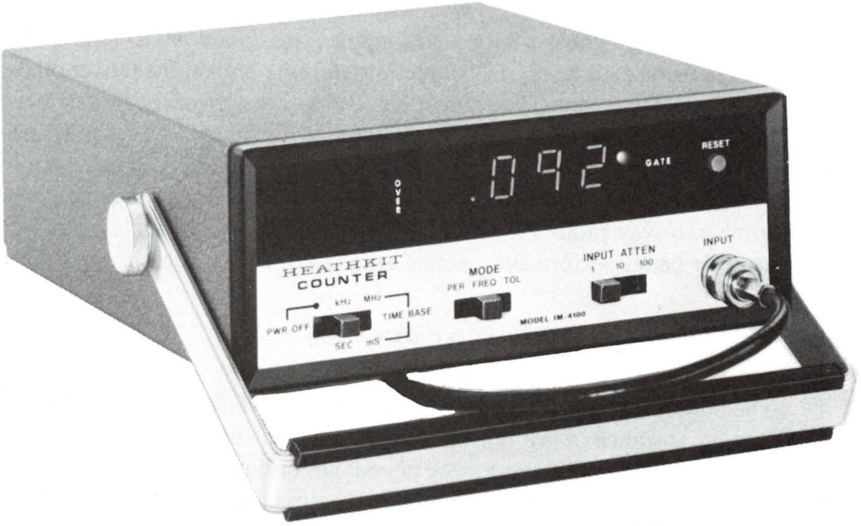


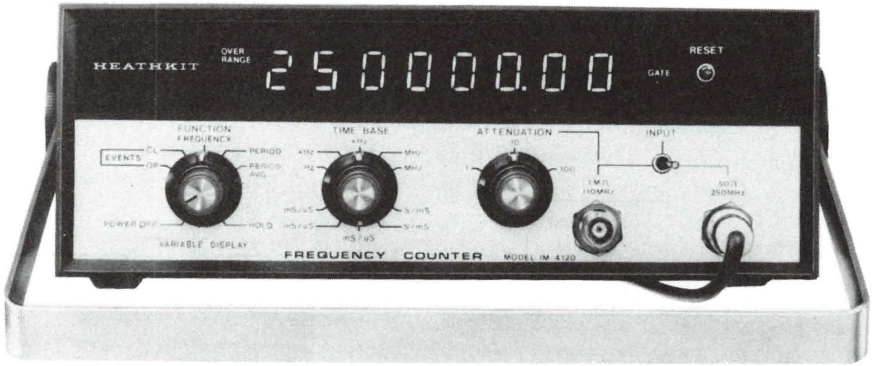
Figure 17-9. Heterodyne frequency meter.



(a)

Figure 17-10. (a) Digital frequency counter. (Courtesy of The Heath Co.)

go back and review appropriate sections of that chapter at this time. Additional commercial examples are given in Chapter 22, "Radio Transmitter Measurements."



(b)

Figure 17-10. *(cont.)* (b) Heath Model IM4120 frequency counter. (Courtesy of The Heath Co.)

17-14 USING FREQUENCY METERS

In the sections to follow we will demonstrate common techniques used to measure frequency. Most of these techniques are found in common usage, even though the use of the digital counter has become predominant over other methods in the past few years. The counter has always been the most convenient for many types of measurement, but until technological spinoffs from the space program reduced IC costs while improving performance, digital counters were too costly for most people. It was not unusual to see the 500-mHz counter that is needed by some two-way radio shops costing over $5000, yet today, models offering the same or better performance can be purchased for considerably less than $1000.

17-15 USING FREQUENCY COUNTERS

The simplest method for checking a transmitter output frequency with a digital counter is shown in Fig. 17-11. A short whip antenna is connected to the input of the counter through a short piece of coaxial, or directly through a coaxial connector mounted on the base of the whip. The transmitter is then keyed on, and the radiated signal is picked up by the whip antenna. If the counter is close enough to the transmitter (less than 500 feet, depending upon power level of the transmitter), sufficient signal will be applied to the input of the counter to cause triggering. The transmitter output frequency is then displayed on the counter readout. Caution is required to measure an unmodulated signal; be duly cautious per the instructions on triggering given in Chapter 6.

Because the signal radiated in this test may interfere with other users of the same or adjacent frequencies, it is a proper test only on the citizens band and in the amateur (ham) radio bands. Even there, it is legal only after listening on the frequency to make sure that nobody is using it.

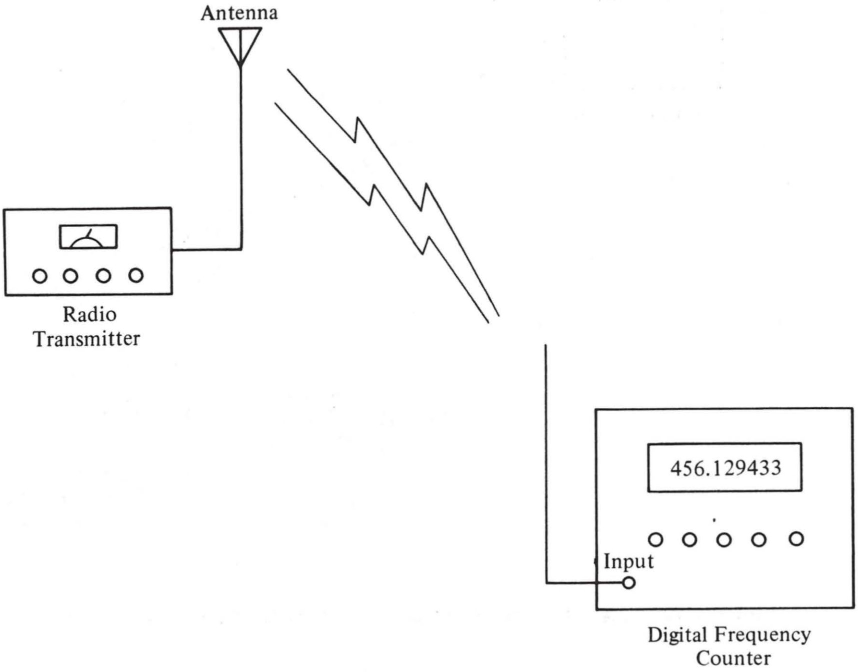


Figure 17-11. "On the air" method of frequency counting (*Note: this technique is illegal except on the citizens band and when Monitoring on-the-air broadcasting stations*.)

To prevent interference with normal communications traffic, it is necessary to measure the frequency by keying the transmitter into a nonradiating dummy load. One version of this technique is shown in Fig. 17-12. Here the transmitter output power is absorbed by the dummy load. A coaxial "tee" connector and a 50-ohm attenuator feeds a portion of the signal to the frequency counter. The attenuator is used to reduce the transmitter's power level to a point where it will not harm the counter input stages. If insufficient attenuation is used for the power level applied, then damage to the counter *will* result.

CAUTION: Never feed the output of a transmitter directly into a frequency counter unless instructed to do so by the manufacturer.

Another legal method for measuring transmitter output frequency is shown in Fig. 17-13. This technique employs a "throughline" RF sampler that picks off a small portion of the transmitter's output signal and routes it to the digital frequency counter. Some models of counter are equipped with a throughline, and they are the *only* types that may be connected directly to the transmitter's output.

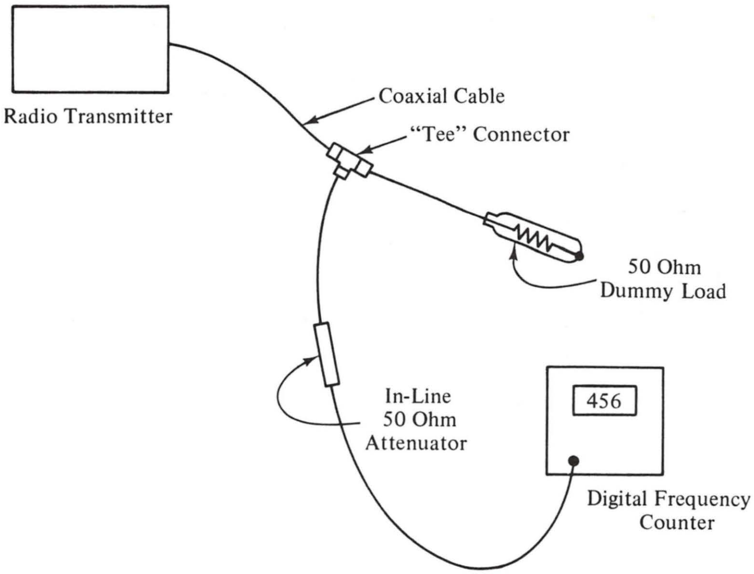


Figure 17-12. Correct method for measuring transmitter frequency with a counter.

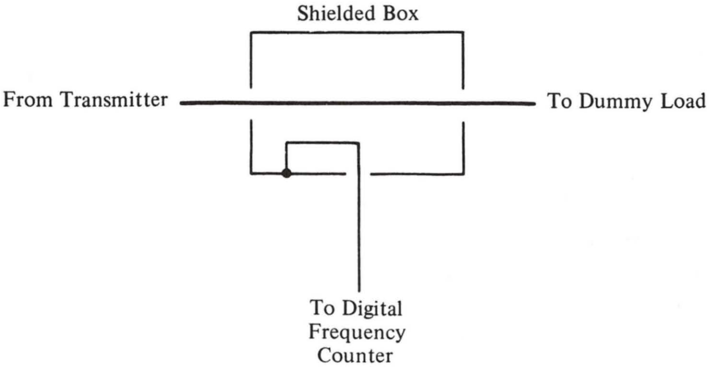


Figure 17-13. Shielded pick-off box.

17-16 TRANSFER OSCILLATORS

A *transfer oscillator* is a stable, precisely calibrated, low-frequency oscillator that is loosely coupled to a communications receiver through a small value capacitor or a wire "gimmick"—a few turns of insulated wire wrapped around the receiver antenna lead.

There are several different ways to use a transfer oscillator in making frequency measurements. It is assumed that we know the approximate frequency of the transmitter from information given as to its channel of operation.

In one transfer oscillator technique the oscillator is a precision device operating on a subharmonic of the transmitter frequency. The frequency of the transfer oscillator is adjusted until it zero-beats with the unknown signal in the receiver. In this case, a harmonic of the transfer oscillator beats against the unknown signal. It is, therefore, important to know which harmonic is being used. We can then multiply the transfer oscillator frequency by the harmonic factor to obtain the actual transmitter frequency.

Example 17-2

It is necessary to measure the frequency of an AM broadcast station that operates on an assigned frequency of 680 kHz. A precision square-wave generator is used as the transfer oscillator. The frequency resolution of the generator can be accurately interpolated to 0.1 Hz. It is found that the zero-beat occurs between the broadcast signal and the generator when the generator frequency is 136,001.1 Hz. What is the transmitter's frequency?

SOLUTION

We know that 136 kHz 1/5 of 680 kHz, so we must be using the fifth harmonic of the generator to zero-beat the transmitter signal. Therefore,

$$F_x = 5 \times 136{,}001.1 \text{ Hz}$$
$$F_x = 680{,}005.5 \text{ Hz} = \mathbf{680.0055\ kHz}$$

The error is 680,005.5 Hz − 680,000 Hz, or 5.5 Hz, well within the ±20-Hz specification for AM broadcast transmitters.

The transfer oscillator technique described in Example 17-2 is limited to relatively low frequencies, where it is easy to obtain variable frequency oscillators that have believable dial calibration. At higher frequencies, the usefulness of the transfer oscillator deteriorates markedly, because the transfer oscillators are less well-behaved. At these frequencies it might be more useful to use the method detailed in Fig. 17-14.

In this method the output signal from the transfer oscillator is fed through a switch. When the switch is in position 1, the oscillator is adjusted to zero-beat with the unknown signal. After this has been done, the switch is transferred to position 2, and a frequency meter is used to read the frequency of the transfer oscillator. Again, we may be using a subharmonic of the actual frequency. Again, we must know which harmonic is being used. When the frequency is high, it is even more necessary to know exactly which harmonic is being used. If

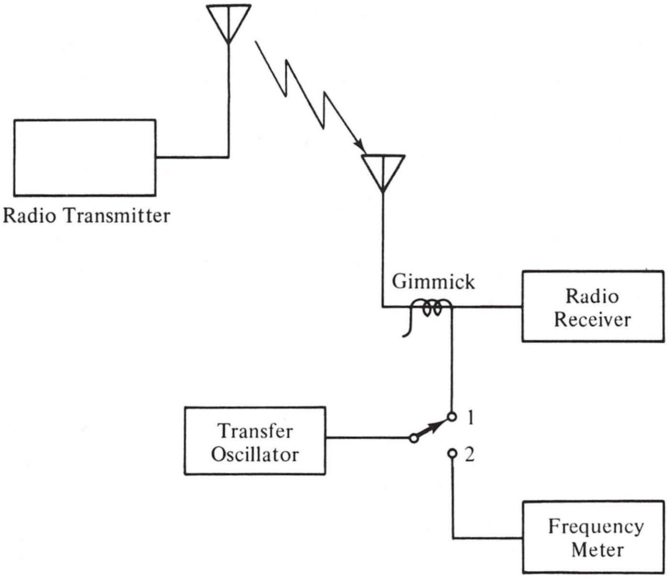


Figure 17-14. Transfer oscillator method of measuring frequency.

extremely low-order (i.e., 1/10 or less) harmonics are used, then it is possible for two harmonics to be simultaneously audible in the receiver output.

This technique is also useful when the transfer oscillator operates on the same frequency as the transmitter, but the transmitter is at a distant location from the measurement site.

17-17 CRYSTAL MARKER OSCILLATORS

A technique that is related to the transfer oscillator is shown in Fig. 17-15. Instead of a transfer oscillator, which is a *variable* frequency oscillator, we use a precision quartz crystal oscillator that operates on a single, standard frequency, or several harmonically related frequencies. This type of oscillator uses a piezoelectric quartz crystal as the frequency control element. In general, crystal oscillators are more accurate and more stable than either R-C or L-C types. The primary requirement for a crystal oscillator, apart from accuracy and stability, is that the output waveform be rich in harmonics. A square wave meets this requirement, but any nonsinusoidal waveform will probably be useful up to the fifteenth or so harmonic, or even further.

Examples of crystal oscillators often used as marker oscillators are shown in Fig. 17-16. In both cases, the frequency-determining network is crystal-controlled. Also, the frequency dividers for both cases are the same, the only

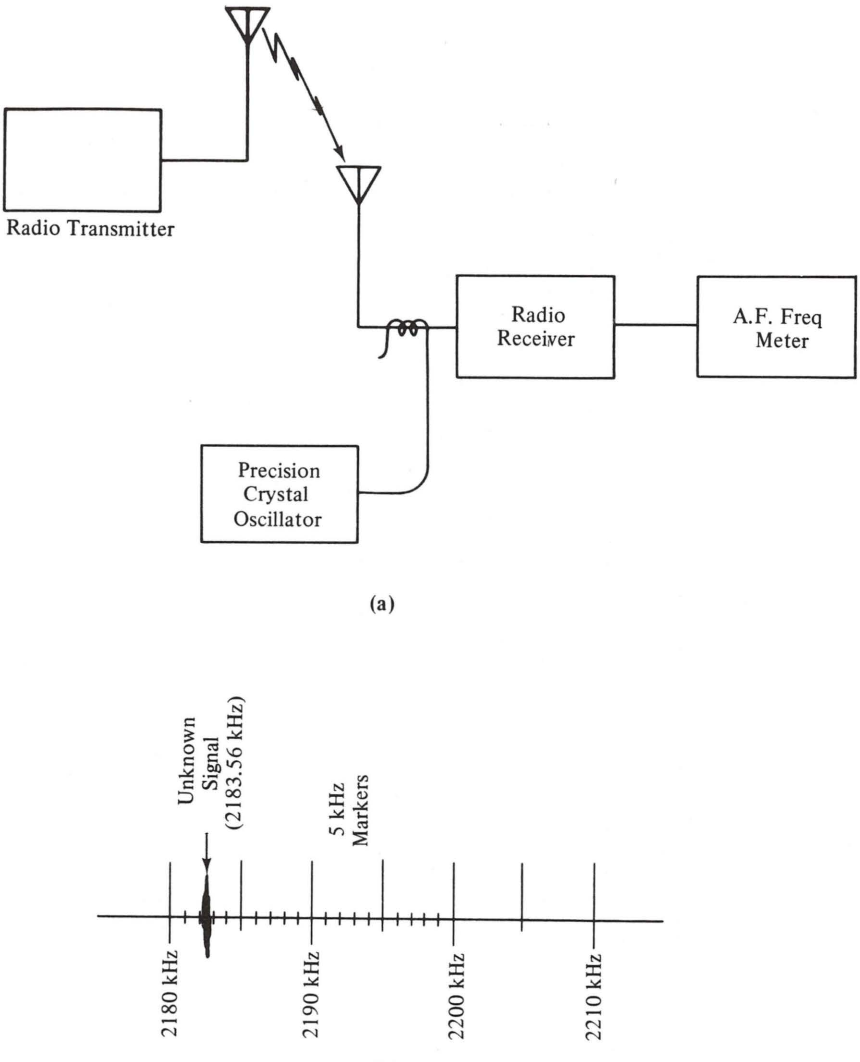
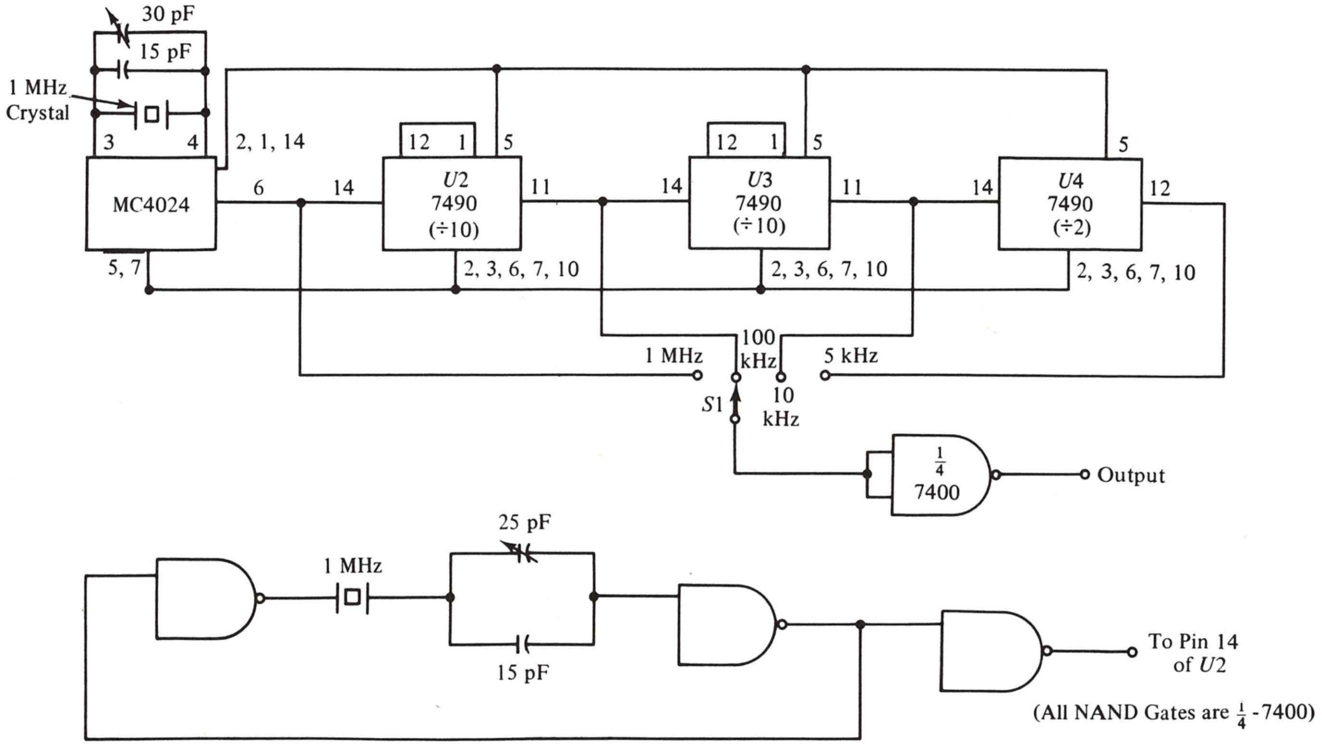


Figure 17-15. (a) Crystal oscillator method of frequency measurement. (b) Receiver dial points during measurement.

difference being the oscillator IC device. The frequency division consists of IC decade dividers that produce outputs of 1 mHz, 100 kHz, 10 kHz, and a divide-by-2 that produces 5-kHz output.

Two different oscillator circuits are shown in Fig. 17-16. The main circuit is preferred by the author. It uses a special IC device made by Motorola, called



(All NAND Gates are $\frac{1}{4}$-7400)

Figure 17-16. Crystal marker oscillator using digital ICs for frequency division. Inset: alternate oscillator circuit.

the MC4024 dual VCO. The variable capacitor across the crystal is used to set the oscillator frequency.

An alternate crystal oscillator is shown in Fig. 17-16(b). This circuit uses three TTL NAND gates that are part of a single type 7400 IC device. Again, a variable trimmer capacitor is used to fine-tune the oscillator frequency.

The crystal oscillators shown in Fig. 17-16 can be used for moderately precise applications. Where high precision is required, however, a commercially built crystal oscillator can be used instead. A wide variety of such oscillators are sold. Some are in a fancy cabinet complete with a regulated power supply, whereas others are less fancy, being built into a metal-cased, but epoxy-encapsulated package.

Two designs are prevalent. In the older models, the crystal and often some other components are housed inside a temperature-controlled oven that usually operates at a temperature of 75°C. The later approach, and the one favored in many modern designs, is to temperature-compensate the circuit so that changes in temperature are not reflected as changes in output frequency. Such oscillators are called *temperature-compensated crystal oscillators,* or TCXO.

The output of a crystal oscillator is loosely coupled to the receiver's antenna circuitry through a low-value capacitor, or a wire gimmick. This superimposes the marker signal on top of the other signals, without overloading the receiver input.

There are two or more different ways to use the crystal oscillator markers to make frequency measurements with the aid of a communications receiver. The first, and most crude, is to use the marker oscillator to temporarily calibrate the receiver's dial over a small range near the unknown frequency. Most higher-quality communications receivers have a movable dial pointer to make easier this type of recalibration. For example, let us say that we want to calibrate a receiver to exactly 7.0 mHz. If we connect a 1-mHz crystal oscillator to the antenna input circuitry, then the marker signals will appear every 1 mHz up the band. We then locate the marker that falls very close to the 7-mHz point, and then move the dial cursor until it lies exactly over the 7.0 mark. For those frequencies close to 7 mHz, the dial will be relatively accurate. But one limitation to this method is that receiver dials rarely have better than 100-Hz resolution, and even that level is accomplished only in the higher-priced models.

Some authorities feel that the best technique for using the crystal marker oscillator is to allow it to heterodyne against the unknown signal,. and then to measure the frequency of the audio beat note. An audio frequency meter is used for this purpose in Fig. 17-15, and it is usually connected to the receiver's audio output jack or terminals. For very close resolution, it may be connected directly to the output of the AM detector so that a near-dc beat note can be obtained.

The receiver is calibrated in the same manner as in the previous example. It is usually best to start with a higher frequency marker and then work down, so that you can home in on the unknown frequency and thereby avoid mistakes.

Figure 17-15(b) shows a simulated receiver dial. Also shown are the markers from the crystal oscillator, and the location on the dial of the unknown signal. The generator is first adjusted to produce a 1-mHz output. It is found that the unknown falls between 2 mHz and 3 mHz. The marker generator is then readjusted to produce a 100-kHz signal, and it is found that the marker lies between 2100 kHz and 2200 kHz. Next, the frequency of the marker is set to 10 kHz, and it is found that the unknown is between 2180 kHz and 2190 kHz. Since the unknown, in this case, falls closer to one of the 10-kHz markers, it is unnecessary to go to any lower marker frequency.

Recall that we are seeking a *beat note*. If the unknown is anywhere near the midpoint between two adjacent markers, there will be *two* beat notes of similar frequency. This ambiguity is cleared up only by going to a lower frequency marker in order to gain some additional frequency resolution. In our present case, however, the unknown is close to one of the 10-kHz points, so no ambiguity will exist.

The beat note produced in the audio output is found to be exactly 3562 Hz, as measured on an audio frequency meter. This means that the unknown frequency is located at a point 3562 Hz away from the 2180-kHz marker. Since we are able to observe on the receiver dial that the unknown is located on the high side of the 2180-kHz point, we know that the actual frequency is 2180 kHz + 3562 Hz, or 2,183,562 Hz. Once again, it is desirable, and considered good engineering practice, to measure frequency three times, and then average the results. Tune the receiver away from the unknown, and then back again to the point where you believe the dial to be tuned exactly on the unknown frequency. Do this several times, to take advantage of the possibility that errors will "add out."

The crystal marker oscillator may be calibrated by zero-beating its output against WWV/WWVH, WWVB, or a local atomic frequency standard or other precise source. If aural zero-beats are used, be aware that the typical error, or uncertainty, is 30 Hz (more in observers with poor hearing acuity). This error is *additive*. If you have a 30-Hz uncertainty when you beat the oscillator to a primary frequency standard, and then another 30-Hz error when you beat the oscillator signal against an unknown, then the actual uncertainty may be as high as 30 + 30, or 60 Hz. And that is the best that will be possible for an aural zero-beat immediately after calibration of the oscillator; it will deteriorate further as the oscillator ages. Use nonaural zero-beat methods for any but trivial cases.

17-18 SUMMARY

1. Low-precision audio frequency measurements can be made by using Lissajous patterns on an oscilloscope screen.
2. Several types of frequency standard are used: atomic clocks, gas cell devices, crystal oscillators, and the NBS radio broadcast stations (WWV, WWVH, and WWVB).

3. Two basic categories of frequency meter are used in common practice: heterodyne and counter.
4. Three types of heterodyne frequency meter are used: direct, transfer oscillator, and crystal oscillator.
5. It is not legal to measure a transmitter's frequency off the air except in the citizens band, amateur radio bands, and standard broadcast bands.

17-19 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

17-20 QUESTIONS

1. Define *frequency*.
2. The *units* commonly employed in electronics to measure frequency are ____________, ____________, and ____________.
3. Define *period* and *duration* as applied to electrical waveforms. Note the *difference* between these terms.
4. Units commonly employed in electronics to measure time are ____________, ____________, ____________, and ____________.
5. Give the "power-of-10" relationship between the units expressed in (a) Question 2 and (b) Question 4.
6. *Period* can be measured roughly on an ____________ and more precisely on a ____________ counter.
7. Describe a method whereby audio range frequencies can be measured to two or three decimal points of a single cycle.
8. List three devices that can be used to make rough frequency measurements in the high-frequency range.
9. List the operating frequencies of WWV/WWVH.
10. List the operating frequency of WWVB.
11. Draw the circuits for two types of absorption wavemeter.
12. List three types of absorption wavemeter.
13. Describe in your own words the phenomenon on which the grid dip oscillator depends.
14. List two types of transmission line device capable of rough frequency measurements in the VHF, UHF, and microwave ranges.
15. What accuracy is possible if the devices of Questions 14 are used?
16. WWV frequencies are maintained to an accuracy of ____________.

17. Audio tones broadcast by WWV and WWVH are __________ Hz, __________ Hz, and __________ Hz, and each tone has a duration of __________ seconds.
18. A female voice is heard making announcements on a 15-mHz standard time station. The call letters are __________.
19. When using WWV or WWVH to zero-beat a digital counter's time base oscillator, it is prudent to use the __________ frequency that produces a strong local signal.
20. The best accuracy possible when one is adjusting an oscillator frequency using the aural zero-beat technique is ± __________ Hz.
21. List two types of atomic frequency standard.
22. List a common gas cell type of frequency standard.
23. An atomic beam clock frequency standard has a potential accuracy of __________.
24. List two basic forms of frequency meter.
25. What is the approximate resonant frequency of a cesium beam? (Use four significant figures.)
26. In the U.S. only three types of transmitter may have their frequency checked off the air: __________, __________, and __________.
27. __________ and __________ oscillators are examples of heterodyne frequency meters.
28. Only counters equipped for __________ monitoring can be connected directly to the transmitter. All others must be protected by connecting them to the transmitter through an __________.
29. The __________ technique of frequency measurement uses a local subharmonic oscillator to zero-beat the unknown frequency. The output of the local oscillator can be measured by an external frequency meter.
30. A frequency measurement device consists of an inductor, capacitor, and lamp in series, and is called an __________ __________.
31. A __________ frequency meter zero-beats a calibrated signal source with an unknown signal.

17-21 PROBLEMS

1. What is the frequency of a repetitive waveform that has a period of 856 nanoseconds?
2. What is the period of a 2.2250-mHz signal?
3. An oscilloscope is used to measure the *period* of a sine wave. The oscilloscope time base is set to 1 μs/div, and the sine wave is observed to occupy 8.6 horizontal divisions. What is the period?
4. What is the frequency of the signal measured in Problem 3?

5. A period counter reads 4.60211 microseconds. What is the frequency of the applied signal?
6. A 1:1 Lissajous pattern rotates in a direction that indicates that the unknown frequency is above the known frequency. Find the unknown frequency if the known frequency is 91 Hz, and the pattern rotates at 12 times per minute.
7. A 100-kHz crystal oscillator is zero-beat against 10 mHz WWV, the aural method being used. The resultant oscillator frequency will be between __________ Hz and __________ Hz.
8. A WWVB comparator receiver is adjusted so that $N_1 = 64$ and $N_2 = 1024$ when the output meter indicates the zero-beat condition. What is the unknown frequency?
9. A pair of Lecher wires is used to measure the operating frequency of a radio transmitter. The first null is noted at a distance of 28.2 cm, and the second null is at a distance of 56.45 cm. What is the probable frequency of operation?
10. A crystal oscillator and a communications receiver are used to measure a frequency near 9 mHz. It is found that a 1098-Hz beat note is heard and that the unknown frequency is below the 9-mHz marker. Calculate the unknown frequency.
11. A transmitter is required to be within ±0.004 percent of its assigned frequency. Express this in ppm.
12. A transmitter must be operated on an actual frequency of 27,085 kHz ±0.005 percent. The actual measured frequency is found to be 27,086,023 Hz. Is the transmitter within the specified tolerance?

17-22 REFERENCES

1. *NBS Publication 236: "NBS Frequency & Time Broadcast Services,"* The National Bureau of Standards, Washington, D.C.
2. *Reference Data for Radio Engineers,* 5th ed. Indianapolis, Ind: ITT-Howard W. Sams, Inc., 1968.
3. Joseph J. Carr, *Elements of Electronic Communications.* Reston, Va.: Reston Publishing Co., 1977.

chapter 18

MEASUREMENTS ON UNTUNED AMPLIFIERS

18-1 OBJECTIVES

1. To learn to make gain and power output measurements on untuned amplifiers.
2. To learn to make harmonic distortion measurements on amplifiers.
3. To learn intermodulation distortion measurement techniques.
4. To learn the operation of principal audio test instruments.

18-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside of it, and then look for the answer as you read the text.

1. Total harmonic distortion (THD) is measured with a low distortion oscillator, an ac voltmeter, and a ____________ filter.
2. Define voltage gain in your own words.
3. Write the transfer function for a voltage amplifier.
4. In frequency response tests of a voltage amplifier we plot ____________ vs ____________.

18-3 UNTUNED AMPLIFIERS

Almost all electronic circuits or instruments contain untuned amplifiers. These amplifiers are generally thought of as *audio* amplifiers, but that is an erroneous assumption. The audio amplifier is merely one type of untuned amplifier. A video amplifier is also untuned, but its frequency response may extend to 50 or 100 mHz, certainly far above "audio." At the other end of the spectrum, an ECG or EEG amplifier used in medical electronic instruments is certainly untuned, but its frequency response only extends to 100 Hz on the high end, and as low as 0.05 Hz (decidedly subaudio) on the other end of the range.

18-4 CONVERTING TO DECIBEL NOTATION

Many measurements require power, voltage, or current data, but are best expressed in a unit called *decibels,* denoted dB. The dB is a *ratio* (and nothing more), so it is well suited for expression of amplifier gain, attenuator loss, frequency response, and anything else where it is required for us to *compare* two levels. For electrical power we express decibels as

$$\text{dB} = 10 \log_{10} \frac{P1}{P2} \qquad (18\text{-}1)$$

where dB = the gain or loss in decibels
$P1$ and $P2$ = the two power levels
$\log_{10}$ refers to the base-10 logarithms

Example 18-1

Calculate the power gain of an amplifier if 100 mW applied to the input produces an output power of 14 watts.

SOLUTION

$$\text{dB} = 10 \log_{10} (P1/P2) \qquad (18\text{-}1)$$
$$\text{dB} = (10) \log_{10} (14/0.100)$$
$$\text{dB} = (10) \log_{10} (140)$$
$$\text{dB} = (10)(2.15) = \mathbf{21.5\ decibels}$$

Similarly, we can express voltage and current gains in decibels usings Eqs. (18-2) and (18-3), respectively.

$$\text{dB} = 20 \log_{10} (E1/E2) \qquad (18\text{-}2)$$
$$\text{dB} = 20 \log_{10} (E1/E2) \qquad (18\text{-}3)$$

Example 18-2

Calculate the voltage gain of an amplifier if 100 mV causes an output potential of 14 volts.

SOLUTION

$$\text{dB} = 20 \log_{10} (E1/E2) \qquad (18\text{-}2)$$
$$\text{dB} = 20 \log_{10} (14/0.1)$$
$$\text{dB} = (20) \log_{10} (140)$$
$$\text{dB} = (20)(2.15) = \mathbf{42.9\ decibels}$$

Examples 18-1 and 18-2 showed the use of decibel notation for power and voltage *gains;* that is, the numerator was made the higher number. But what about *losses?* A convention that has been adopted is to place the *output* figure in the numerator, so that the lesser number is in the numerator in the case of a loss. This will result in exactly the same number of decibels, but the fact that it is a loss shows up because the result is negative. If the "output/input" protocol is followed, then it is always possible to distinguish losses and gains in making system calculations.

Example 18-3

An attenuator drops a 10-volt signal to 50 mV. Calculate the loss in decibels.

SOLUTION

$$\text{dB} = 20 \log_{10} (E1/E2) \qquad (18\text{-}2)$$
$$\text{dB} = 20 \log_{10} (0.05\ \text{V}/10\ \text{V})$$
$$\text{dB} = 20 \log_{10} (0.005)$$
$$\text{dB} = (20)(-2.30) = \mathbf{-46\ decibels}$$

Note that the *loss* caused the result of the decibel equation to be negative, where it had been *positive* for *gains*. The same relationship is found also in current and power ratios.

18-5 SPECIAL DB-BASED SCALES

Decibel scales are merely *ratios* between two sets of numbers. It is, however, sometimes convenient to reference all measurements within a system, or even an entire industry, to a standard level that is usually referred to as "0 dB." If this is done, then we may simply add up all the gains and losses, and then note how the result compares with the amount of signal required to do the job at hand. We may then use the data to predict signal levels and certain aspects of performance and to make changes in the design to take care of any problems.

In several areas of electronics we see these special dB scales. Some of the most common are given below:

1. *dBm* (used in RF measurements). Defined 0 dBm as 1 milliwatt dissipated in a 50-ohm load.
2. *VU* (audio *volume units*). The VU scale is a decibel-based scale in which 0 VU is defined as 1 milliwatt dissipated in a 600-ohm load at 1000 Hz.
3. *dB* (obsolete). Defines 0 dB as 6 milliwatts dissipated in a 500-ohm load at 1000 Hz.

4. *dBmv* (TV coaxial cable antenna systems). Defines 0 dBmv as 1 mV across 75 ohms. This is the only one of the four that is not defined in terms of power.

The user of a special dB-based scale may never be required to convert a measurement back to the basic units, i.e., milliwatts or millivolts. The dB data will usually suffice. If a conversion is necessary, then it is a simple matter to substitute in the standard power or voltage level, as given above, and then solve the appropriate decibel equation for the unknown, such as it may be.

18-6 VOLTAGE GAIN MEASUREMENT

The gain of a voltage amplifier is defined by the amplifier transfer function:

$$A_v = \frac{E_{out}}{E_{in}} \tag{18-4}$$

where A_v = the voltage gain (dimensionless)
E_{out} = the output signal voltage
E_{in} = the input signal voltage

We can *measure* the gain of an ampilifier by measuring the input signal voltage and the output signal voltage, and then taking the ratio expressed by Eq. (18-4). Figure 18-1 shows the equipment configuration used to perform many of the tests that we will discuss in this chapter. We are assuming that ac (i.e., audio) amplifiers are being tested, but the same procedures also hold true for dc amplifiers if a dc signal source and dc voltmeters are substituted for their ac counterparts.

Procedure

1. Adjust the attenuation, oscillator output level, and amplifier gain controls for a convenient output level. (In some cases, the input voltage, total harmonic distortion point, or output level may be further specified.)
2. Measure the input and output signal voltages.
3. Make the gain calculation, using Eq. (18-4).
4. If dB gain figures are desired, then use a modified version of Eq. (18-2): dB = $20 \log_{10}(A_v)$.

The voltage gain measurement is usually made with the amplifier gain at maximum (if it is variable) and the input signal level at some point well below the level where output clipping is noted on the oscilloscope, or the specified THD level is reached.

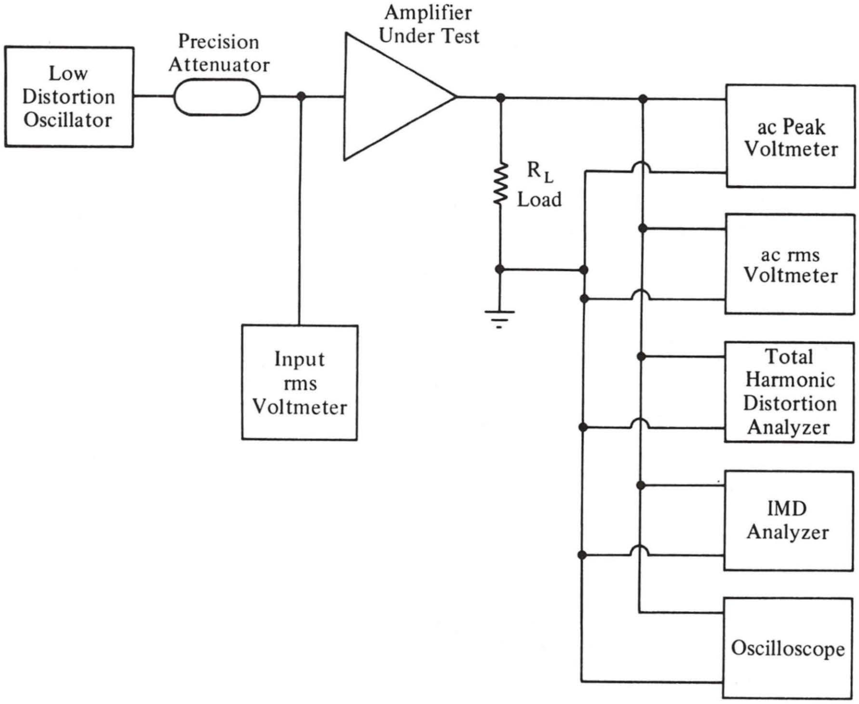


Figure 18-1. Test equipment setup for performing most audio tests and measurements.

18-7 OUTPUT POWER MEASUREMENT

Hi-fi output power was one measurement that was often used to confuse the buyer with unrealistically high numbers, rather than to convey anything really coherent about the product being sold. We used to mutter something about "taffy being distributed . . ." when sales people talked of hi-fi amplifier power levels. There were seemingly innumerable abuses of measurement procedure that produced amplifier ratings in "watts," but were essentially meaningless.

It would seem that such a "basic" measurement would be very cut and dried. But we often found the following types of "wattage" ratings: (a) rms watts at 1 percent THD, (b) rms watts at 5 percent THD, (c) peak watts, (d) peak-to-peak watts (!?!), and (e) something called "music power." Of these, only the rms watts rating accurately reflects the ability of the amplifier to do work, that is, move a loudspeaker cone. But note well that most amplifiers produce considerably more output power if the maximum permissible THD is 5 percent, than they will if only 1 percent THD is permitted.

Let us consider these various types of "power measurement" with respect to a proper measurement of rms watts at 1 percent THD. We have a hypothetical power amplifier connected to an 8-ohm resistive load in a circuit such as Fig. 18-1. It is nominally a 25-watt rms amplifier. At a THD of 1 percent we find that it produces an rms output potential of 14 volts ac.

1. Rms watts at 1 percent THD
   Equation: $P = (E_{\text{rms}})^2/R$
   Solution: $P = (14)^2/8 = 24.5$ watts
2. Peak watts
   Equation: $P = (E_{\text{p}})^2/R$
   Solution: $P = (1.414 \times 14)^2/8 = 49$ watts
3. Peak-to-peak watts
   Equation: $P = (2.83E_{\text{rms}})^2/R$
   Solution: $P = (2.83 \times 14)^2/8 = 196$ watts
4. "Music power"
   Equation: $P = 2 \times$ rms power
   Solution: $P = (2)(14)^2/8 = 49$ watts

Note the wide range of numerical results (and all of these have been advertised!). But only watts rms at 1 percent THD is particularly meaningful to the user, although some public address applications can tolerate 5 percent THD, so there one is able to use the watts rms at 5 percent THD rating. Perhaps you can see why some people used to refer to "watts IYL" (If You're Lucky) to describe audio amplifier power specifications.

Procedure

1. Adjust the amplifier for maximum gain and increase the level from the oscillator until either (a) clipping begins (and then back off until clipping disappears), or (b) until the THD increases to 1 percent (or the manufacturer's specified level).
2. Measure the rms voltage across the load resistor.
3. Calculate the rms power from the expression $P = (E_{\text{rms}})^2/R_{\text{L}}$.
4. In most cases, this measurement is performed at 1000 Hz, or some other frequency in the 200–2000-Hz range specified by the manufacturer. Some feel, however, that a somewhat more stringent test is to measure the power at 1000 Hz and the upper and lower -3 dB points. The results of the three measurements are then averaged.

Audio power measurements can be deceiving, even when the rms rating is used. For example, in a *stereo* amplifier, the power supply must be capable of delivering sufficient power to produce maximum output from both channels. Some manufacturers will specify the output power with *both channels driven.*

This is to take into account the regulation of the power supply, which must be able to handle twice the power load than when only one channel is driven. The output power from either channel should be nearly the same under both conditions; if the "both channels driven" power is significantly less than the power measured with only one channel driven, then suspect a weak power supply.

18-8 INPUT SENSITIVITY

The sensitivity rating of an amplifier tells us how much signal is required to drive the amplifier to a specified output level. Two methods seem generally applied: drive level required for (a) full output power, and (b) 1 watt output.

Procedure

1. Adjust the amplifier gain to maximum.
2. Increase the input signal level to produce the specified output level.
3. Measure the rms input voltage.
4. The sensitivity is usually written in the form "6 mV for 1 watt output."

18-9 FREQUENCY RESPONSE

The *frequency response* of an amplifier refers to its ability to amplify signals of different frequencies. Ideally, a perfect amplifier will amplify all frequencies from dc to daylight equally. But in practical amplifiers this ideal is not met.

The response plot of an amplifier (Fig. 18-2) graphs *gain vs frequency* on semilogarithmic paper. The relative gain is not actually measured, but is inferred from the output voltage; the input signal voltage is kept a constant.

The gain, then, is expressed in terms of output volts, which are usually converted to decibel form. The 0-dB reference level will be specified as the output voltage at a given reference frequency, usually 400 or 1000 Hz. The reference frequency must be far enough removed from the upper or lower −3 dB points that the response is essentially flat.

In most cases, the frequency response data are taken manually by adjusting the oscillator frequency and then reestablishing the input voltage to the amplifier using the attenuator. The output voltage and frequency are then recorded.

In some cases, however, modern voltage controlled (VCO) signal generators are used to *sweep* the spectrum. If a sawtooth sweep voltage is used for both the VCO and the oscilloscope time base, then the amplitude of the amplifier output signal (applied to the oscilloscope Y input) represents the frequency response data.

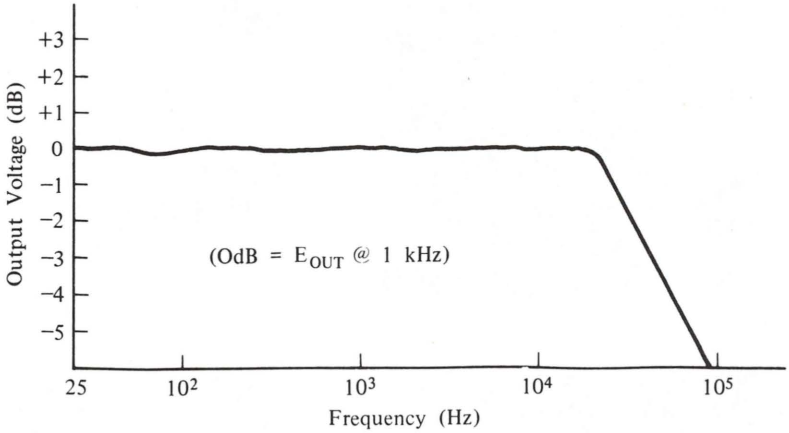


Figure 18-2. Frequency response curve.

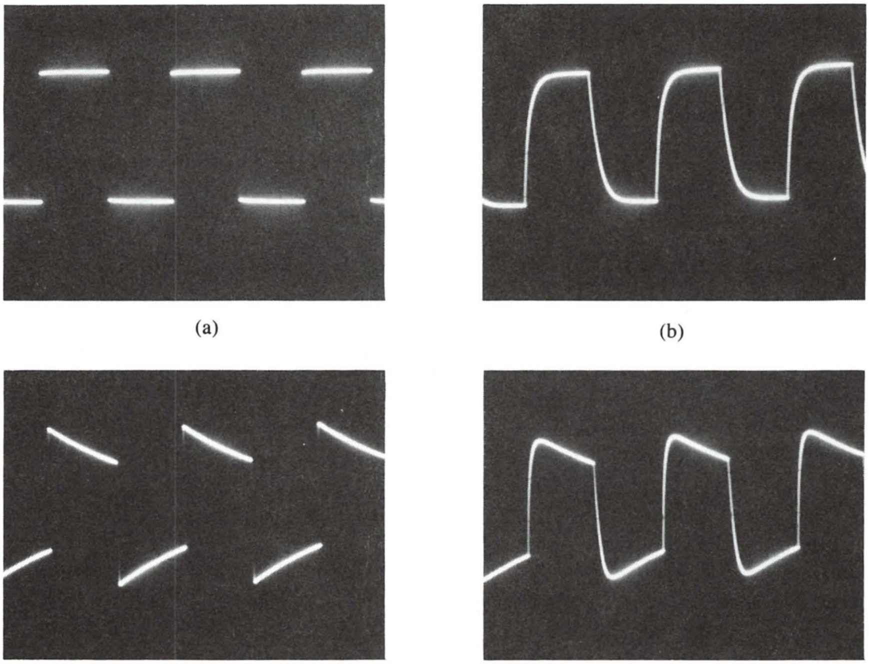
18-10 SQUARE WAVE TESTING

If a square wave input signal to the amplifier is used, then we may gain a qualitative indication of frequency response by examining the output waveform. Although quantitative data are lacking, this test gives a quick method for making a qualitative evaluation of the amplifier's response.

The square wave test is possible because of the nature of square waves. If you perform a mathematical analysis of a square wave to find its Fourier series, then you will find that it is made up of a fundamental sine wave at the frequency (some say repetition rate) of the square wave, plus an infinite number of *odd* sine wave harmonics. Each successive harmonic has less amplitude than the previous harmonic. Real square waves, i.e., those that are produced in a signal generator, have up to approximately the one-thousandth harmonic in significant amplitude.

Figure 18-3 shows various types of square wave response. In Fig. 18-3(a) we see the response of a very wide band amplifier, which is essentially the input square wave unaffected. In very wide band amplifiers only slight rounding of the upper corners may be expected, and this may not be visible on the oscilloscope trace. Figure 18-3(b) shows the results of passing a square wave through an amplifier that has a much more limited frequency response. Note the severe rounding of the corners. It is the *upper* harmonics that contribute to the rise time and "sharpness" of the waveform corners.

Poor low-frequency response produces a trace like that shown in Fig. 18-3(c). In this case, the leading edges of both positive and negative excursions of the pulse are "overpeaked," and the flat portion is tilted.



(a) (b) (c) (d)

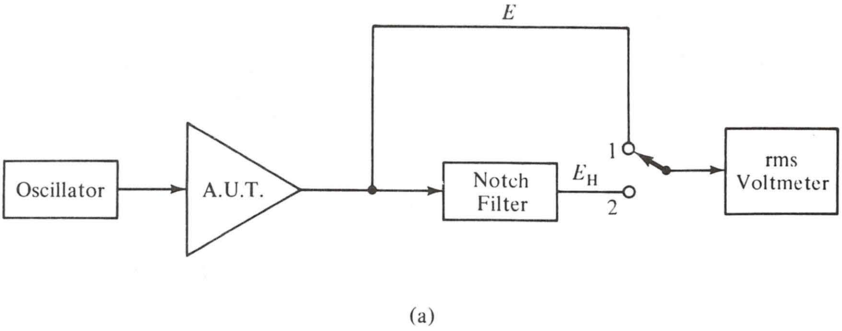
Figure 18-3 (a) Normal square wave. (b) Loss of high frequencies. (c) Loss of low frequencies. (d) Loss of both high and low frequencies.

18-11 TOTAL HARMONIC DISTORTION (THD)

All waveforms (and in fact all continuous mathematical functions) can be described in terms of a Fourier series of sine and cosines, and in most cases they bear harmonic relationships to the fundamental sine wave frequency. The specific harmonics that are present and their relative amplitudes and phases determine the actual shape of the waveform. Only the sine wave is pure; all other waveforms have harmonic components.

When an amplifier distorts a waveform, it is creating harmonic components that have not existed before. The measure of this type of distortion is the percentage of *total harmonic distortion* (THD).

Total harmonic distortion is measured with a harmonic distortion analyzer, blocked out in Fig. 18-4(a). The oscillator supplying the input signal to the amplifier under test must have a very pure (that is, harmonic-free) sine wave output signal. The THD analyzer does not care where the distortion originates, and will measure the total harmonic distortion, regardless of the source.



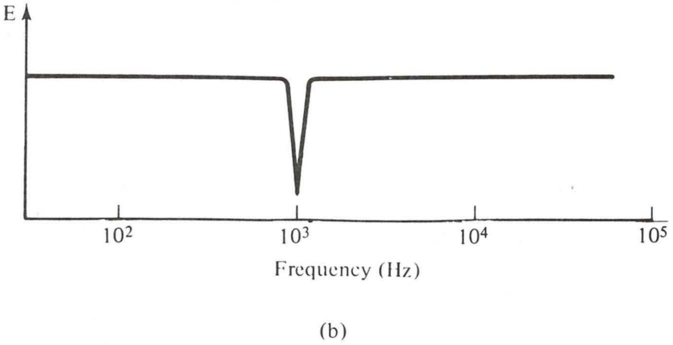


Figure 18-4. (a) Total harmonic distortion analyzer. (b) Notch filter frequency response.

An rms ac voltmeter is used to measure the distortion. We first measure the rms voltage of the amplifier output potential E, and then measure the rms voltage due to harmonics E_h. This voltage is measured by passing the signal through a *notch filter* that is adjusted to pass all frequencies except the oscillator's output frequency, i.e., the fundamental frequency. The response of a typical notch filter is shown in Fig. 18-4(b). The percentage THD is expressed by

$$\text{THD} = \frac{E_h}{E} \times 100\% \tag{18-5}$$

where THD = the percentage of total harmonic distortion
E_h = the rms output voltage measured after the notch filter
E = the rms output voltage measured before the notch filter

Example 18-4

Calculate the percentage of THD if the amplifier output voltage is 7 V rms, and the notch filter output voltage is 56 mV (the notch filter is tuned to the oscillator frequency, precisely).

SOLUTION

$$\mathrm{THD} = (E_{\mathrm{h}}/E) \times 100\%$$
$$\mathrm{THD} = (0.056\ \mathrm{V}/7\ \mathrm{V}) \times 100\%$$
$$\mathrm{THD} = (8 \times 10^{-3}) \times 100\% = \mathbf{0.8}\%$$

Note that THD performance for any given amplifier varies with output power level. An amplifier in which THD is measured at 1 watt output, for example, appears far more linear than one in which the THD was measured at maximum rated output power. An example of a commercial THD analyzer, the Sound Technology, Inc. Model 1700B, is shown in Fig. 18-4(c).

(c)

Figure 18-4. *(cont.)* (c) Sound Technology Model 1700B distortion analyzer. (Courtesy of Sound Technology, Inc.)

18-12 INTERMODULATION DISTORTION (IMD)

When two signals are combined together in a linear network, they will both pass through the system without affecting each other. Figure 18-5(a) shows the result of mixing a high-frequency signal (i.e., 3–4 kHz) with a higher amplitude low-frequency signal (i.e., 40- 100 Hz) in a linear network. If the output signal of the network were examined on a spectrum analyzer, only the two original frequencies would be present.

If these same signals were passed through a nonlinear network (i.e., an amplifier that produces distortion), then mixing or "heterodyning" will result, because the lower frequency will tend to modulate the higher frequency. If $F1$ and $F2$ are the two original frequencies, then the output spectrum would contain the following (at least): $F1$, $F2$, $(F1 + F2)$, and $(F1 - F2)$. There might also be present harmonics of these signals, plus the sum and difference frequencies caused by the harmonics.

If there is any significant nonlinearity in the circuit to cause distortion, then the effect on the two tones will be as in Fig. 18-5(b), the familiar amplitude modulation waveform. The percentage of intermodulation distortion is given by

$$\text{IMD} = \frac{b - a}{a} \times 100\% \qquad (18\text{-}6)$$

where IMD is the intermodulation distortion in percent
b, a refer to feature in Fig. 18-5(b)

A simple IMD analyzer (Fig. 18-6) can be made by merely rectifying the amplifier output signal. A standardized test signal is used in this circuit. The higher frequency must be at least 50 times greater than the lower frequency. For a 60-Hz low frequency, then, the high frequency must be at least 3000 Hz. The two signals are combined in a linear resistor network, usually a Wheatstone bridge in which all four arms are equal.

The test circuit uses a high-pass filter to reject the low-frequency component of the output signal, and pass only the high-frequency component plus the modulation products (i.e., sum and difference frequencies), which lie near the high-frequency signal. A low-pass filter is used to average the rectifier output to remove any residual 3000-Hz signal. The remaining signal is a dc voltage that represents the 60-Hz component that has been modulated onto the 3000-Hz "carrier." Since the original high-pass filter removed the main low-frequency component, it does not affect this reading. The output voltage, then, is proportional to the percentage of modulation, or, in this case, intermodulation.

In untuned amplifiers, i.e., audio amplifiers, IM distortion is also sometimes called cross modulation. When one is dealing with radio receivers, however, this terminology may be misleading. In receiver systems, intermodulation and cross modulation have very similar meanings—but there is a slight difference. In speaking of untuned amplifiers, however, both terms are used, even though the term *intermodulation* is preferred.

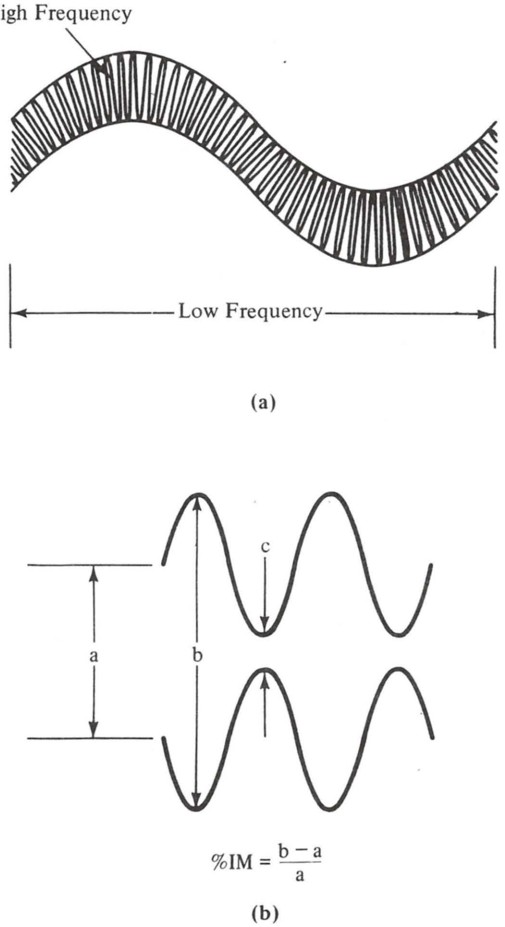


Figure 18-5. (a) Linear mixing of two tones. (b) Nonlinear mixing of two tones produces AM waveform.

18-13 OTHER PARAMETERS

In this section we will discuss certain other aspects of untuned amplifiers. Many of the tests discussed so far in this chapter are particularly interesting when one is testing hi-fi and other audio amplifiers. But tests in this section are used primarily in other types of amplifier, as might be found in scientific, industrial, medical, or engineering instruments. Some of these tests are specific to *differential* amplifiers, whereas others are applicable to amplifiers in general. We will use the operational amplifier for our examples, but this is not to imply that the tests are not valid on other types as well.

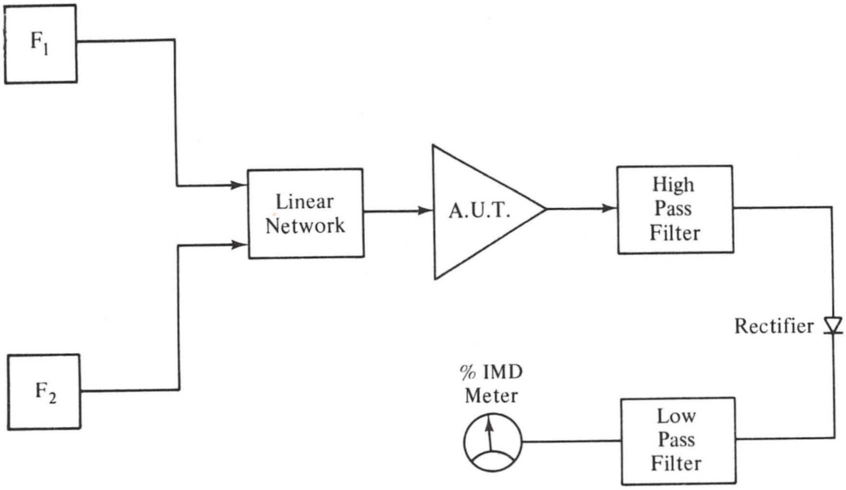


Figure 18-6. Test equipment needed for IMD measurement.

18-14 SLEW RATE (S_r)

The *slew rate* S_r of an amplifier is the measure of its maximum *change* of output *voltage* (i.e., dE/dt), while the amplifier is delivering its full rated output current. Figure 18-7 shows the test configuration used in this test.

The rated output current I_m and the maximum output voltage E_m are determined from the amplifier's specifications, and the value of R_L is determined by Ohm's law, E_m/I_m.

The unity gain noninverting follower circuit is used in Fig. 18-7(a) because, in operational amplifiers especially, this configuration produces the worst case results. An inverting follower, or noninverting follower *with* gain, can also be used, but the results cannot be expected to apply to all cases.

The input signal E_{in} is adjusted to severely overdrive the amplifier in both positive and negative directions [see Fig. 18-7(b)]. The *slope* of the output voltage waveform, as viewed on an oscilloscope, represents the slew rate.

$$S_r = \frac{\Delta E_o}{\Delta t} \qquad (18\text{-}6)$$

The units of slew rate are in *volts per unit of time*. In actual practice, the units commonly seen are *microvolts per second* or *volts per microsecond*. Although both representations are valid, it can lead to some confusion unless you are aware of which is being used.

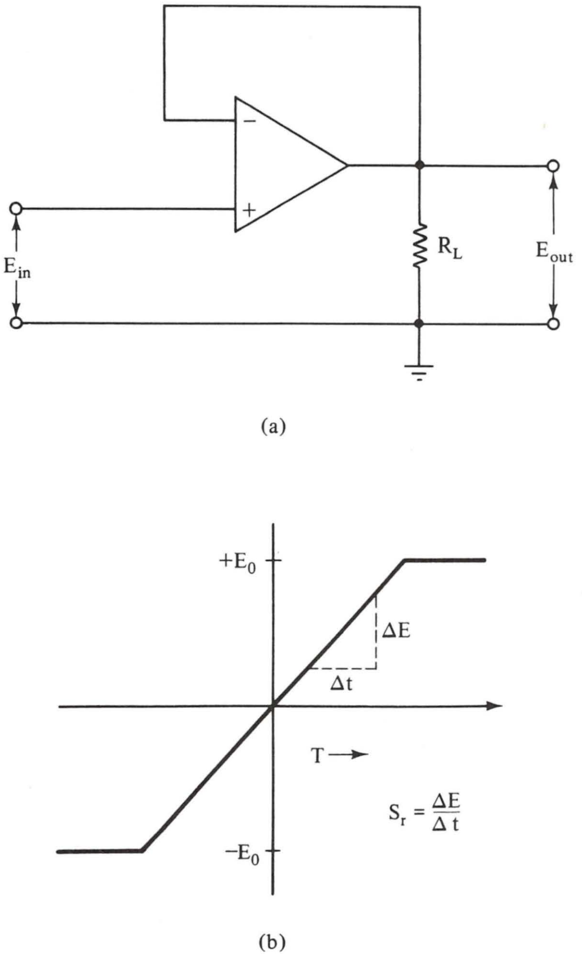


Figure 18-7. (a) Noninverting, unity gain follower used in test circuit. (b) Definition of slew rate.

18-15 FULL POWER BANDWIDTH

The *full power bandwidth* of an amplifier is the frequency at which open-loop voltage gain (A_{vol}) drops to unity, while the amplifier is delivering its full rated output power. This frequency will usually be somewhat less than the gain-bandwidth product for small signals.

It is rarely feasible to measure this parameter in an open-loop configuration, because noise sources ordinarily present in all circuits will predominate in the output signal. Fortunately, we may approximate this parameter using a gain follower circuit, provided that the closed-loop voltage gain is set to something in the ×100 to ×500 range. The full power bandwidth at these gains will approximate that of A_{vol} to a negligible error.

The input signal amplitude is set to a level that drives the output to deliver full rated power to an external load resistor R_L. The frequency is then increased, with the input voltage maintained constant, until the output voltage E_0 is equal to the input voltage E_{in}, i.e.,

$$A_v = E_0/E_{\text{in}} = 1$$

18-16 INPUT OFFSET VOLTAGE

The *input offset voltage* is defined as the potential across the inverting and noninverting inputs required to bring the output potential to zero at a time when E_{in} is also zero. We would normally expect E_0 to be zero any time E_{in} is zero, but certain circuit and internal IC defects often cause a small erroneous offset voltage.

The measurement circuit is shown in Fig. 18-8. It is an inverting follower with a gain of ×500 to ×1000. The "input" end of resistor $R1$ is grounded to ensure that E_{in} is zero. Then a high resolution, precision, dc power supply is applied directly across the amplifier inputs. The supply is adjusted until E_0 is zero. The voltage E that reduces E_0 to zero is the input offset voltage of the amplifier.

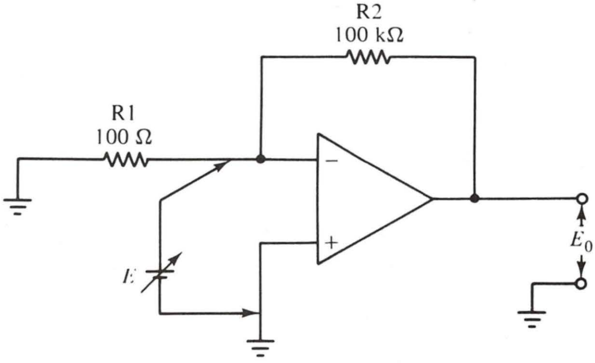


Figure 18-8. Test setup for measuring input offset voltage.

18-17 COMMON MODE REJECTION RATIO

Common mode rejection is the property of a *differential* amplifier that rejects input signals that are *common* to *both* inputs. The *common mode rejection ratio* (CMRR) is expressed by

$$\mathrm{CMRR} = \frac{A_{\mathrm{v(d)}}}{A_{\mathrm{v(cm)}}} \qquad (18\text{-}7)$$

$$\mathrm{CMRR} = \frac{(\mathrm{E}_0/E_{\mathrm{in}})}{(E_0/E_{\mathrm{cm}})} \qquad (18\text{-}8)$$

$$\mathrm{CMRR} = \left.\frac{E_{\mathrm{cm}}}{E_{\mathrm{in}}}\right|_{E_0 = constant} \qquad (18\text{-}9)$$

A convenient differential voltage is selected, and both the input voltage E_{in} and the output voltage that it produces, E_0, are measured precisely. Next, the two inputs are tied together, and used as a single, common mode input. The input signal voltage is increased until the same output voltage as was obtained before is at the output. Equation (18-9) can then be used to calculate CMRR.

18-18 SUMMARY

1. Voltage gain is the ratio of output voltage over input voltage.
2. The *rms* power measurement is more closely related to the work that an audio power amplifier can do than are some of the other "power" measurements sometimes used.
3. The total harmonic distortion (THD) is found by measuring the rms output voltage with and without the fundamental present.
4. Intermodulation distortion is measured by noting the degree of modulation of a high-frequency signal by a low-frequency tone.

18-19 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

18-20 QUESTIONS

1. Define in your own words what is meant by a "decibel scale."
2. Power or voltage gain is represented by ____________ dB, while a gain or loss is represented by ____________ dB.

3. Write the transfer equation if a voltage amplifier has an input signal $E1$ and an output signal $E2$.
4. Discuss in your own words the different power measurement techniques used on hi-fi amplifiers.
5. What is meant by "input sensitivity"?
6. Frequency response is a measure of ____________ vs ____________.
7. A square wave test gives us a ____________ view of an amplifier's frequency response.
8. Define in your own words "total harmonic distortion."
9. Describe in your own words a technique for measuring THD. (Use a diagram if you want.)
10. Describe in your own words a method for measuring frequency response data. (Use a diagram if you need it.)
11. Describe in your own words a method for measuring ac voltage gain (i.e., at 1000 Hz).
12. Define in your own words "intermodulation distortion."
13. Draw a block diagram of an IMD analyzer.
14. Define "slew rate" in your own words.
15. Define "full power bandwidth" in your own words.
16. Define "input offset voltage" in your own words.
17. Draw circuits for measuring (a) slew rate and (b) full power bandwidth—without looking back in the text.

18-21 PROBLEMS

1. An amplifier produces 6 V output when 1.3 mV is applied to the input. Calculate the power gain in both scaler and dB.
2. An RF amplifier produces 650 watts output when a 6-watt signal is applied to the input. Calculate the power gain in scaler units and dB.
3. Calculate the *loss* in dB if an attenuator reduces a 1-V signal to 1 mV.
4. An RF signal source connected to a matching, nonreactive, load is set to produce −10 dBm. How much power is dissipated in the load?
5. A 600-ohm load is connected to an audio sine wave generator that is set to produce +3 VU output. How much *voltage* (rms) appears across the matched load?
6. In a voltage gain measurement 7 V is produced at the output by a 100-mV signal at the input. Calculate the gain in dB.

chapter 19

MEASUREMENT ON TUNED CIRCUITS

19-1 OBJECTIVES

1. To learn some of the problems encountered in making measurements on tuned RF circuits.
2. To learn how the basic spectrum analyzer operates.
3. To learn some applications of the basic spectrum analyzer.
4. To learn sweep frequency techniques.

19-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. Describe briefly in your own words what a spectrum analyzer does.
2. Name the principal stages in a spectrum analyzer.
3. Spectrum analyzer data are said to be in the ____________ domain.
4. The sweep frequency used to test a high-Q resonant tank circuit must be kept low in order to prevent ____________ in the circuit.

19-3 TUNED CIRCUITS

In this chapter we are going to consider certain techniques and instruments used to measure various parameters in frequency-selective, i.e. tuned, circuits. Although any circuit that has a bandwidth less than "dc-to-daylight" may be considered by some to be frequency selective, we are going to limit the discussion here to circuits of relatively narrow bandwidth. Generally, these circuits will be RF in nature, hence the emphasis on RF circuits.

The reader is referred also to Chapters 20 through 22 for an additional study of this subject as it applies specifically to antennas, receivers, and transmitters, respectively. Although an attempt is made at not presenting the same material more than once, there is, of necessity, some overlap. For the broadest view of the subject of RF measurements, then, you are advised to study Chapters 20–22 in addition to this chapter.

Tuned circuits consist of R, L, and C elements in parallel or series combination. Figure 19-1 shows two forms of tuned circuit, the series resonant [Fig. 19-1(a)] and parallel resonant [Fig. 19-1(b)].

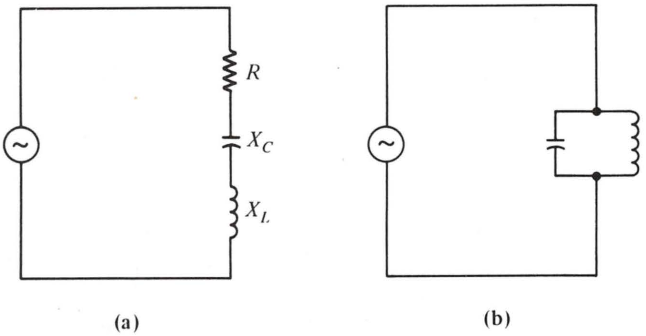


Figure 19-1. (a) Series resonant tank circuit. (b) Parallel resonant tank circuit.

A tuned circuit differs from an untuned circuit because of the property of resonance, which exists in LC circuits when the capacitive and inductive reactances are equal in magnitude; i.e.,

$$X_L = X_C \tag{19-1}$$

$$2\pi FL = \frac{1}{2\pi FC} \tag{19-2}$$

The capacitive and inductive reactances produce opposite effects on ac currents, so when the reactance magnitudes are equal, then the net effect is cancellation of the reactive component. The circuit becomes essentially resistive. The resonant frequency can be determined by solving Eq. (19-2) for F, which results in the expression

$$F = \frac{1}{2\pi LC} \tag{19-3}$$

where F = the resonant frequency in hertz (Hz)
L = the inductance in henrys (H)
C = the capacitance in farads (F)

Example 19-1

Calculate the resonant frequency of a tank circuit consisting of a 100-μH inductor and a 100-pF capacitor.

SOLUTION

$$F = 1/[2\pi(LC)^{½}] \tag{19-3}$$

$$F = \frac{1}{(2)(3.14)[(10^{-4}\text{ H})(10^{-10}\text{ F})]^{½}}$$

$$F = \frac{1}{(2)(3.14)(10^{-14})^{½}} = \mathbf{1.59\ mHz}$$

There is a fundamental difference in the performance of these two types of resonant tank circuit. The graph plotting the *impedance vs frequency* for these circuits is shown in Fig. 19-2. The series-resonant circuit (curve A) presents a high impedance at frequencies removed from resonance, and a low impedance at resonance. The behavior of the parallel resonant circuit is exactly the opposite. The parallel tank circuit offers a high impedance at the resonant frequency, and a low impedance at frequencies removed from resonance.

Note that the response of these circuits falls off gradually from resonance in both cases. The "sharpness" of the fall-off depends upon the *quality factor,* or *figure of merit,* designated by the symbol Q. The Q of a tank circuit is given by

$$Q = \frac{F_r}{\text{B.W.}} \tag{19-4}$$

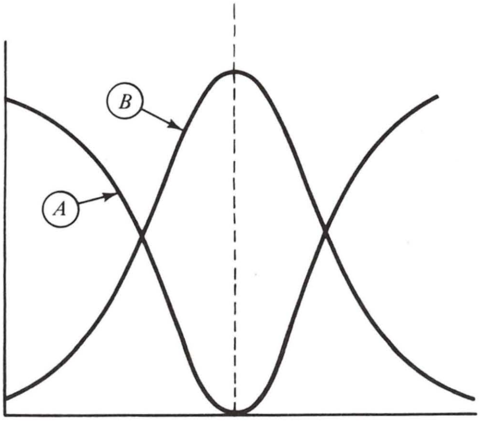


Figure 19-2. (a) Impedance in a series-resonant circuit and (b) impedance in a parallel circuit.

$$Q = \frac{X_L}{R} \tag{19-5}$$

where Q = the quality factor
F_r = the resonant frequency*
B.W. = the bandwidth*
R = the circuit resistance

Figure 19-3 shows the frequency response curves for tank circuits with varieties of Q values. Curve A shows the response given when the Q has a moderate value, while curve B shows the response obtained when the tank circuit has a high Q value. A tank circuit with a very low Q will produce the curve given in C.

The curve for a moderate Q is marked at a point that is 70.7 percent of the maximum point. This is known as the −6-dB, or "half-power" point. Bandwidth in tank circuits is measured with these points as the guidelines. When someone is heard referring to a bandwidth of "some many kilohertz," he is indicating that number of kilohertz between the points where the response falls off 6 dB.

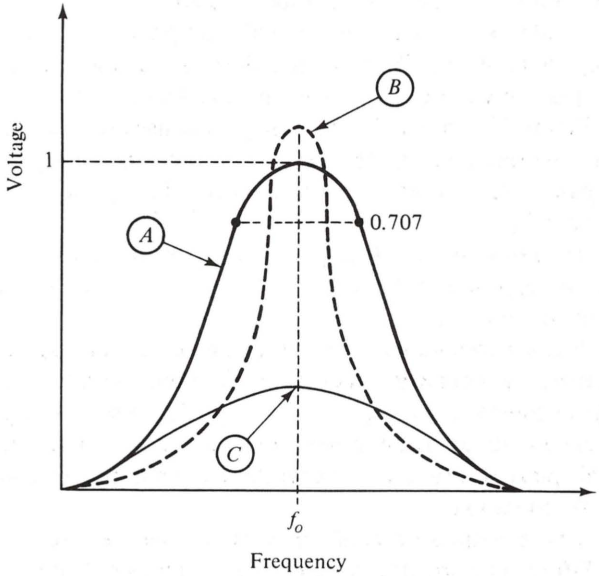


Figure 19-3. (a) Medium-Q, (b) high-Q, (c) low-Q resonant circuit frequency response curves.

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*Any units of frequency may be used, provided that the *same* units are used for both F_r and B.W.

Tuned transformers (Fig. 19-4) are encountered frequently in RF circuits. In many cases both primary and secondary windings of the transformer are resonated by capacitors, whereas in others only one winding is resonated.

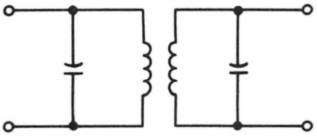


Figure 19-4. Tuned transformer.

The apparent Q of a tuned transformer is affected by the degree of coupling between the primary and secondary tank circuits. Figure 19-5 shows the three basic responses obtained with differing degrees of coupling. Figure 19-5(a) shows the response obtained when *subcritical* coupling is used. In this type of coupling few flux lines from the primary cut the secondary, so only the strongest signals close to the resonant frequency are passed.

The effect of *critical coupling* is the moderate Q curve of Fig. 19-5(b). *Overcritical coupling* produces the low Q "double hump" curve in Fig. 19-5(c). In this last type of coupling most of the primary flux lines cut the secondary. This condition results in the classic response shown.

A filter is a special tuned circuit that passes frequencies in a certain band and rejects all others. The tuned transformer is a case of a *bandpass filter;* that is, it will pass only those frequencies in a band around the resonant frequency.

Figure 19-6 show the frequency responses for several different classes of filters. The response shown in Fig. 19-6(a) is for a *low-pass filter*. This type of filter passes frequencies lower than the cutoff frequency F_c, and rejects frequencies above F_c.

The response for a *high-pass filter* is shown in Fig. 19-6(b). This response is just the opposite of the low-pass filter; it passes frequencies *above* F_c and rejects those below F_c.

If low-pass and high-pass filter elements with slightly overlapping cutoff frequencies are connected in cascade with each other, then we find that we have a *bandpass* filter such as Fig. 19-6(c). A similar response is produced by a simple tank circuit, tuned transformers, and certain other RLC networks. Quartz and ceramic piezoelectric crystal networks also produce the bandpass response under the right conditions.

Two examples of *band reject* filter responses are shown in Figs. 19-6(d) and 19-6(e). The principal difference between these two is in the bandwidth. The response shown in Fig. 19-6(d) rejects a stop band centered about F_r but extending out in both directions quite a bit. The notch filter in Fig. 19-6(e) shows a response in which the stop band is very narrow around the resonant frequency.

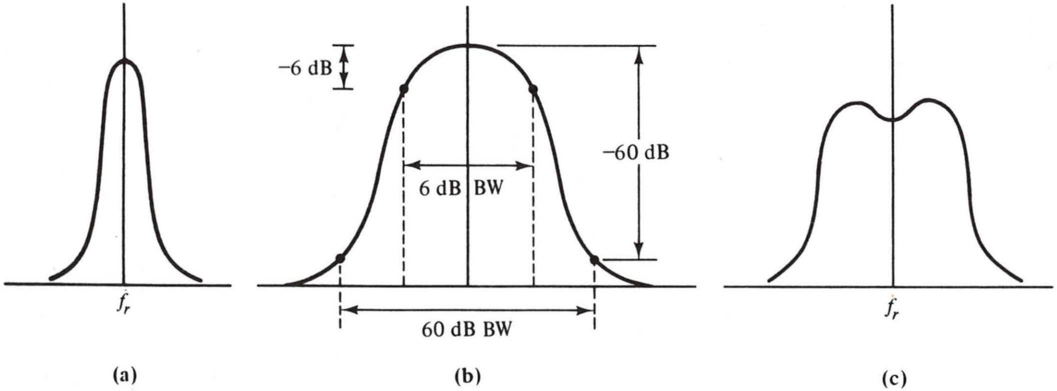


Figure 19-5. (a) Sub-critical coupling, (b) critical coupling and (c) over-critical coupling.

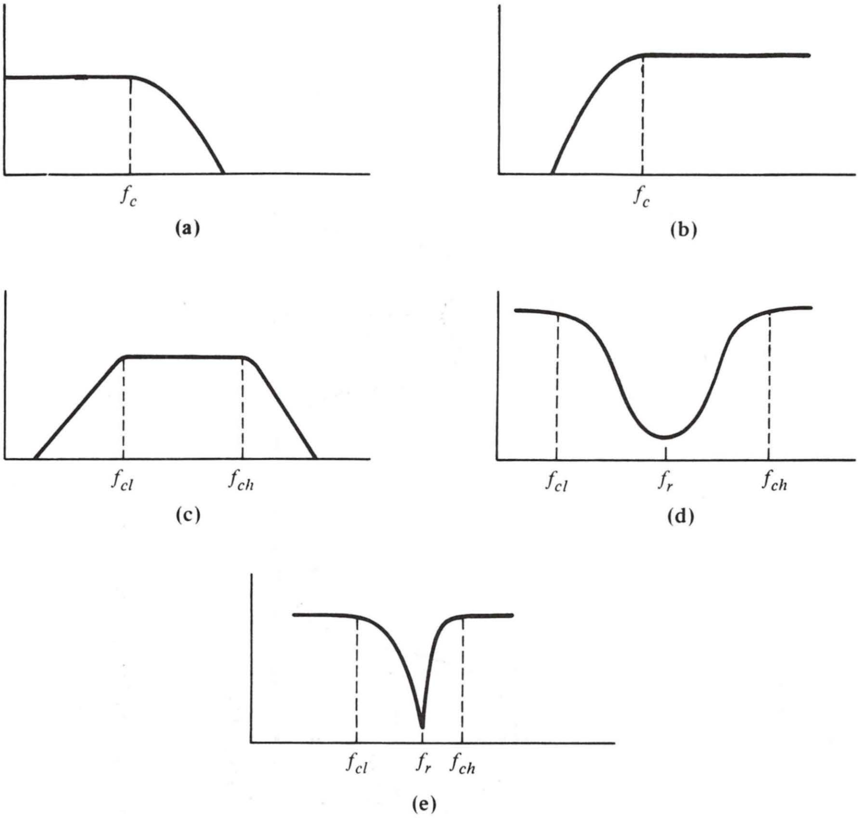


Figure 19-6. (a) Low-pass filter response. (b) High-pass filter response. (c) Bandpass filter response. (d) Band reject filter response. (e) Notch filter response.

19-4 TESTING TUNED CIRCUITS

Tuned circuits tend to be a little more critical of the techniques used than are many untuned circuits. While low frequency tank circuits are relatively well-behaved, the problems increase dramatically with increased frequency.

Several problems will become immediately apparent once you begin to work with tuned circuits, but one of the most critical is loading of the circuit by the test equipment. Suppose, for example, that you want to measure F_r by using a signal generator and an RF voltmeter. The cables to these instruments could easily cause a parallel capacitance across the tank circuit that has a value of several hundred picofarads. This value may be many times the actual tuning capacitance of the tank, but even if it is merely a significant fraction of the tuning capacitance, serious errors in the measurement will result. The solution to this problem is to isolate the tank circuit from the test equipment, whenever possible.

One very gross type of measurement of resonant frequency is to use the dip meter, discussed in Chapter 17. But dip meter calibration is coarse, at best.

A signal generator and RF voltmeter (or oscilloscope) can be used if proper isolation of the tank circuit is provided. If the tank is part of a cascade amplifier, then the normal input/output terminals may be used. Otherwise, it may prove wise to use a "gimmick" to excite the tank, and a very low-capacitance probe for the voltmeter. It sometimes helps to connect the voltmeter or oscilloscope to a low-impedance tap on the tank circuit inductor, or provide a transformer link for the indicator.

The response curve can be measured, if somewhat laboriously, by adjusting the signal generator to various points around F_r, and then manually plotting the data on graph paper. This technique "grows old" very quickly, however, because of the relative difficulty in obtaining the needed data. It is better to use a sweep generator to write the curve onto the oscilloscope CRT screen.

19-5 SWEEP TECHNIQUES

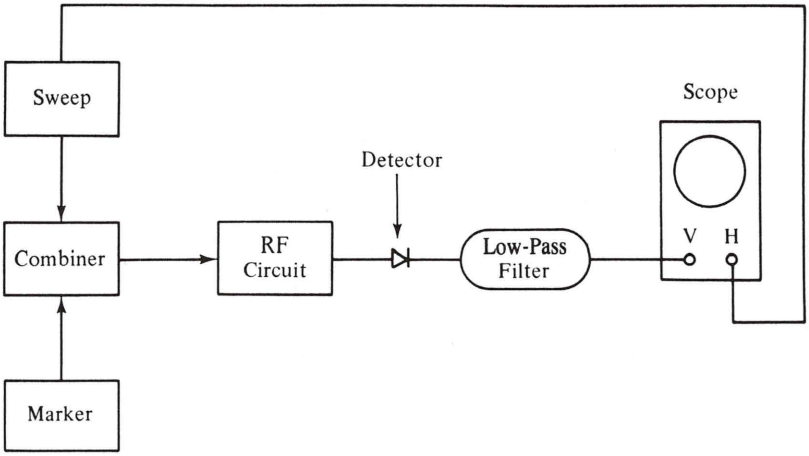
A *sweep generator* (Chapter 9) uses a voltage-controlled oscillator (VCO) driven by a sawtooth or sine wave (in which case it is an FM generator). In either case, we create a varying RF frequency that sweeps back and forth across the frequency set on the dial at a rate determined by the sawtooth or sine wave.

The block diagram of a sweep generator test configuration is shown in Fig. 19-7(a). The sweep generator signal and a signal from a crystal calibrator (i.e., marker) are mixed in a linear network before being applied to the circuit under test. The output of the RF circuit is passed through a detector and low-pass filter to the vertical input of an oscilloscope. The sweep generator will also send either the sweep voltage (i.e., the sawtooth), or a synchronization pulse, to the oscilloscope. In the former case, the sweep generator would be connected to the oscilloscope horizontal channel, whereas in the latter it would be connected to the *sync* or *trigger* input. Figure 19-7(b) shows a typical sweep generator-oscilloscope package; this one is a Tektronix TM-500 series.

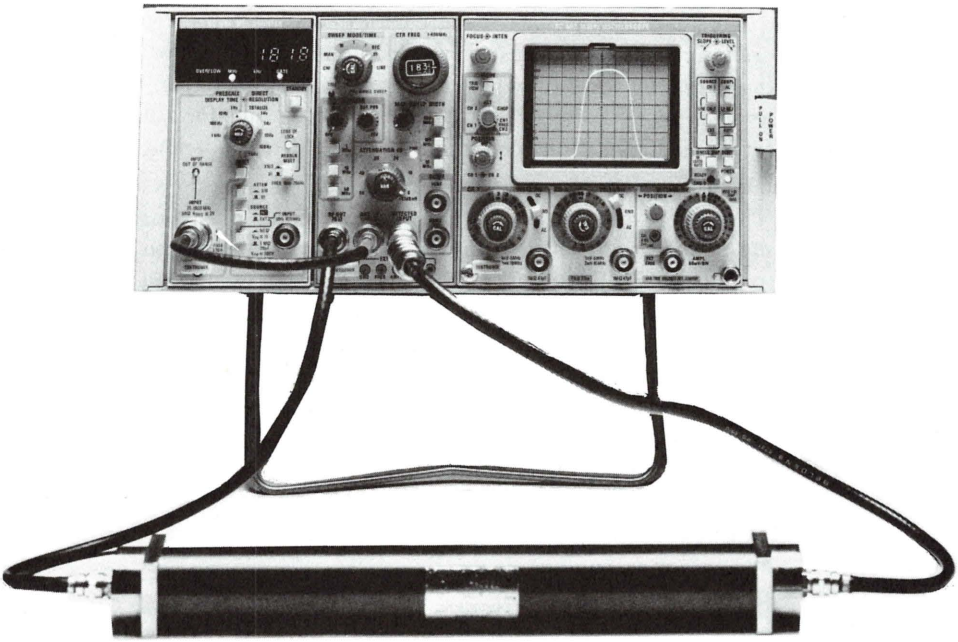
One must be careful when using sweep techniques not to use too high a sweep frequency, or *ringing* of the tank circuit will occur.

19-6 SPECTRUM ANALYZERS

All of the test procedures outlined so far involve the *time domain*. That means that the data are of the *time-vs-frequency* format. But that is only part of the picture, because much interesting information lies in the frequency domain as well.



(a)

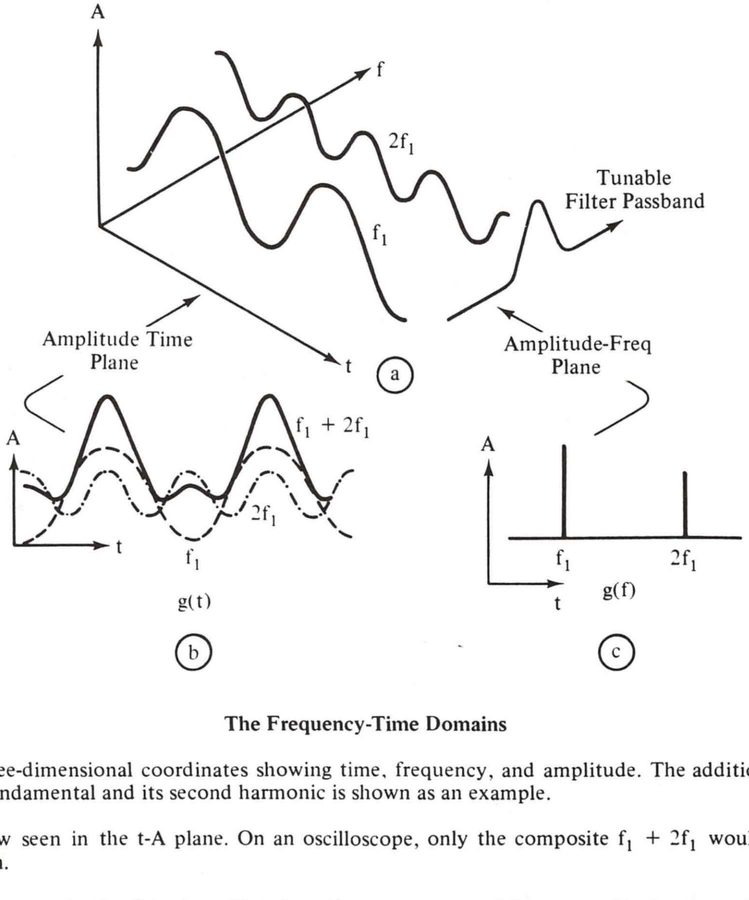


(b)

Figure 19-7. (a) Block diagram of equipment needed for sweep testing. (b) Sweep testing setup. (Courtesy of Tektronix, Inc.)

Figure 19-8 shows the relationship between the time and frequency domains. In this example, we view a waveform that is the summation of a sine wave fundamental and its second harmonic. Figure 19-8(a) shows a three-dimensional coordinate system plotting the *amplitude vs time (A-t)* and *amplitude vs frequency (A-f)* data.

Figure 19-8(b) shows the composite as viewed in the *A-t* plane. The trace produced is the classic waveform for second harmonic interference on a fundamental. But the trace in Fig. 19-8(c) is a little different; it is an *A-f* bar graph showing the relative amplitudes of the two components. A *spectrum analyzer* is an oscilloscope that displays the *A-f* data.



The Frequency-Time Domains

a. Three-dimensional coordinates showing time, frequency, and amplitude. The addition of a fundamental and its second harmonic is shown as an example.

b. View seen in the t-A plane. On an oscilloscope, only the composite $f_1 + 2f_1$ would be seen.

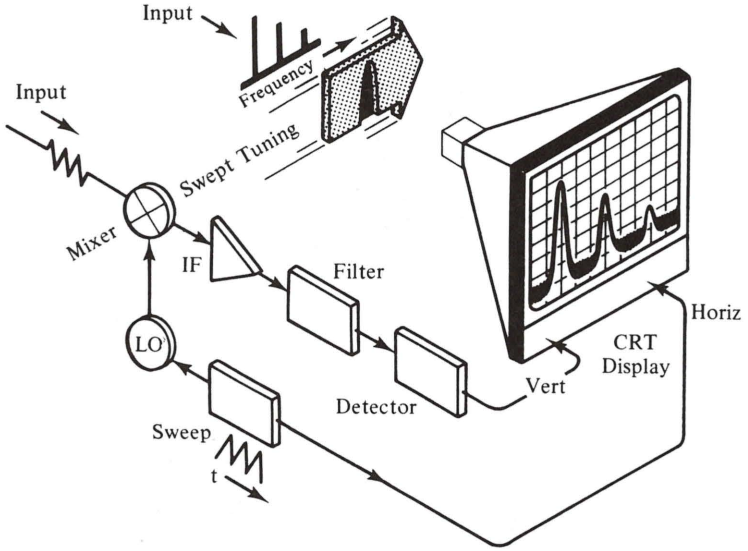
c. View seen in the f-A plane. Note how the components of the composite signal are clearly seen here.

Figure 19-8. (Courtesy of Hewlett-Packard.)

The spectrum analyzer is actually a superheterodyne receiver (Fig. 19-9) in which the local oscillator is a sweep generator. A low-frequency sawtooth wave is applied to both the sweep oscillator and the horizontal deflection plates of the CRT, producing a horizontal deflection that is a function of frequency. The lowest frequency of the sweep oscillator is represented by the left side of the trace, while the highest frequency is represented by the right side. The sweep is left to right.

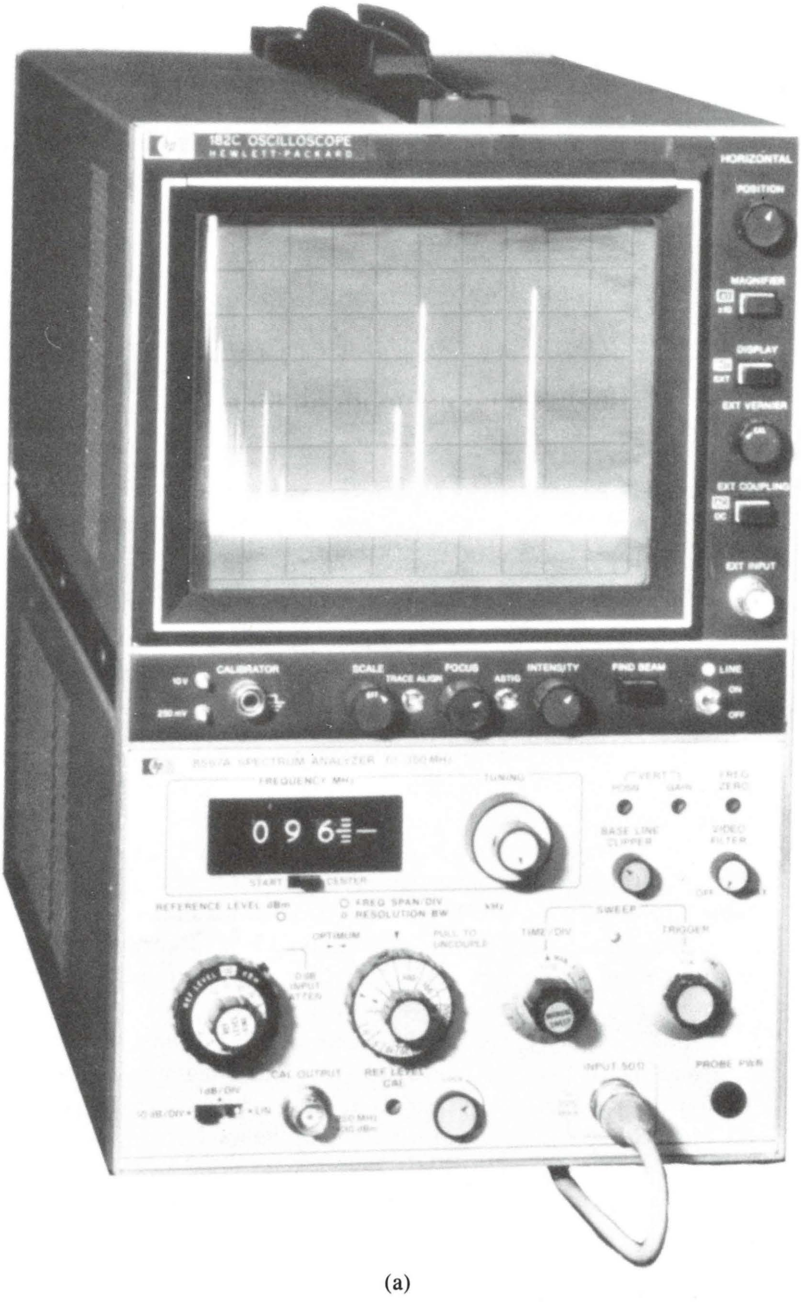
The input signals are mixed with the local oscillator signal, to produce the IF (i.e., difference) signal. The bandwidth of the IF amplifier is relatively narrowband, so the output signal at the detector will have a strength that is proportional to the frequency that the L.O. is converting to the IF at that instant. The display, then, will contain "poles" that represent the amplitudes of the various input frequency components.

Two examples of commercial spectrum analyzers are shown in Fig. 19-10. The example shown in Fig. 19-10(a) is the Hewlett-Packard model 8557A. It has a range of 10 kHz to 350 mHz, and can accommodate signal levels between −117 and +20 dBm. This model has twelve calibrated sweep spans from 5 kHz/div to 20 mHz/div in a 1, 2, 5 sequence. The resolution varies from 1 kHz to 3 mHz, depending upon the range.



The spectrum analyzer is a swept-tuned receiver. The input is mixed with a swept local oscillator, the resultant amplified through an IF strip, and displayed as the vertical deflection on a cathode ray tube. The horizontal deflection of the CRT is proportional to the frequency of the swept LO. For further information see HP Application Note 150.

Figure 19-9. Block diagram of a spectrum analyzer. (Courtesy of Hewlett-Packard.)



(a)

Figure 19-10. (a) Ten-kHz to 350-mHz spectrum analyzer. (Courtesy of Hewlett-Packard.)

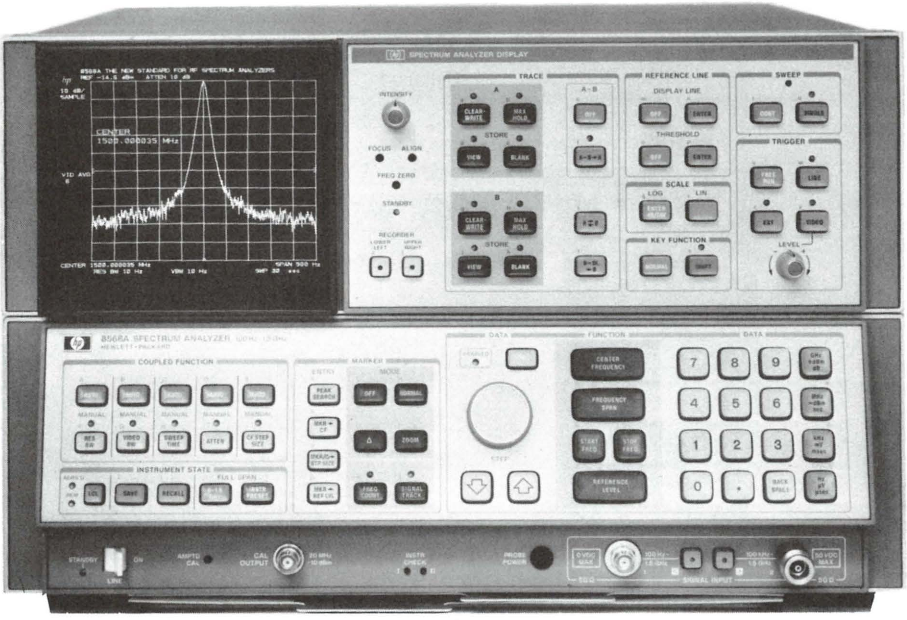

(b)

Figure 19-10. *(cont.)* (b) 100-Hz to 1.5-gHz spectrum analyzer. (Courtesy of Hewlett-Packard.)

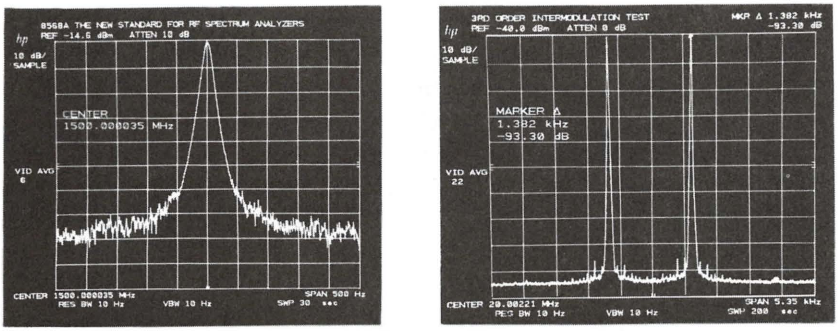
A somewhat wider-range instrument is shown in Fig. 19-10(b); it will cover 100 Hz to 1500 mHz, at input levels from −137 to +30 dBm. Notice that the parameters desired of the measurement are entered via a numeric keyboard.

Examples of spectrum analyzer traces are shown in Fig. 19-11. The trace in Fig. 19-11(a) shows the output spectrum of a 1500-mHz crystal-controlled oscillator. The spectrum analyzer is set for 10-Hz resolution.

The results of a two-tone intermodulation distortion test on a 20-mHz transmitter are shown in Fig. 19-11(b). Note that the output is relatively pure, the third-order IM products being greater than 85 dB down.

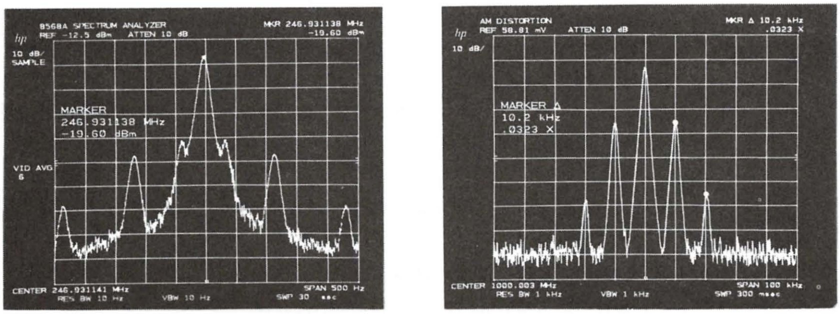
In Fig. 19-11(c) we see the spectrum of a 247-mHz transmitter, with the sidebands produced by a low-frequency modulating signal. The spectrum of a 1000-mHz AM transmitter is shown in Fig. 19-11(d). The sideband contains both the product due to the second harmonic of the modulating signal, as well as the products that are due to the signal itself.

The spectrum analyzer finds application in many areas of technology. Whether it is to check the output of a transmitter, or to find the Fourier series of a signal from a vibration transducer, we may deduce much about the system being tested by using a spectrum analyzer.



1.5 GHz crystal-based signal shown with 10 Hz resolution bandwidth. Note purity of signal and display.

Two-tone test shows 3rd order IM products >85 dB down.



Cavity-tuned VHF oscillator showing 30 Hz, 120 Hz and 240 Hz sidebands.

AM distortion: Marker Δ data indicates frequency and amplitude difference between modulation sideband and its harmonic.

Figure 19-11. (Courtesy of Hewlett-Packard.)

19-7 SUMMARY

1. There are two basic types of resonant tank circuit, series and parallel.
2. Resonance exists in a tank at the frequency where the inductive and capacitive reactances are equal.
3. Q is a measure of how fast the frequency response falls off at frequencies removed from resonance.
4. A sweep oscillator and detector produces an *amplitude-vs-time* trace on an oscilloscope.
5. A spectrum analyzer produces an oscilloscope display of *amplitude vs frequency*.

19-8 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

19-9 QUESTIONS

1. List two types of resonant tank circuit.
2. The series-resonant tank circuit exhibits a ____________ impedance at resonance.
3. The parallel resonant tank circuit exhibits a ____________ impedance at resonance.
4. At resonance, ____________ ____________ and ____________ ____________ are equal, so they cancel, leaving only the resistive component.
5. Define Q in your own words.
6. In tuned transformers, subcritical coupling causes the apparent Q to be ____________.
7. Draw the response curves for (a) critical, (b) subcritical, and (c) overcritical coupling.
8. Draw the typical response curves for (a) high-pass and (b) low-pass filters.
9. What is the principal difference between a stop band and a notch filter?
10. Describe how a sweep generator can be used to measure the frequency response of a circuit.
11. Why must a slow sweep frequency be used in sweep tests?
12. A ____________ ____________ is a device that produces an *amplitude-vs-frequency* trace.

19-10 PROBLEMS

1. Derive the equation for resonance in an LC tank circuit, using the standard reactance equations. Show your work.
2. Calculate the resonant frequency of a tank circuit that consists of a 33-pF capacitor and a 1.2-μH inductor.
3. Find the Q of a 5-mHz tank circuit, if the -6-dB points are 30 kHz apart.
4. Find the Q of a 1000-kHz tank circuit if the coil has an inductance of 450 μH and a 50-ohm series resistance (copper loss of the wire).

chapter 20

ANTENNA AND TRANSMISSION LINE MEASUREMENTS

20-1 OBJECTIVES

1. To learn the various parameters normally of interest in antenna/transmission line systems.
2. To learn what instruments are available for antenna measurements.
3. To learn antenna measurement techniques.
4. To learn how to adjust HF and VHF antennas, using *simple* instrumentation.

20-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, then look for the answer as you read the text.

1. Describe two methods for measuring antenna impedance.
2. What are *standing waves,* and how are they formed in radio transmission lines?
3. How can a *noise bridge* be used to read antenna impedance?
4. What is a *slotted line?* What are *Lecher wires?*
5. *Nodes* and *antinodes* alternate with each other, repeating themselves every ____________ wavelength along the line.

20-3 ANTENNAS

Radio antennas are devices that *radiate* an electromagnetic wave into space when excited by an oscillating electrical current. The radiating effect will be present in some degree at all frequencies of oscillation, but the efficiency is extremely poor except at certain resonant frequencies. At these frequencies the efficiency increases tremendously.

There are many forms of antennas, but we are going to limit this discussion to high frequency (HF) and very high frequency (VHF) dipoles and monopoles. Brief excursions will be made into the UHF and low-microwave region, and to directive arrays, but the reader is advised to consult some of the referenced texts at chapter's end for more detail. This chapter, after all, is designed to discuss some of the measurements made on antennas, not the antennas themselves. Interestingly enough, for those readers who wish to pursue the subject, the amateur radio literature contains many high-quality articles from both theoretical and practical points of view, and are generally more available to the public than are the papers presented in arcane professional journals.

Figure 20-1 shows two fundamental antennas, a half-wavelength dipole [Fig. 20-1(a)] mounted horizontally to the earth's surface, and a vertical monopole that is quarter-wavelength [Fig. 20-1(b)]. These antennas are similar in many ways, but a principal difference is in the nature of the field that is radiated by each. Both will radiate an electromagnetic field in which electric and magnetic field components radiate outward at right angles to each other, forming a plane. But the horizontal antenna radiates a signal in which the *electric* field is *horizontal* to the earth's surface, whereas the vertical antenna radiates a signal in which the electric field is vertical to the earth's surface; i.e., it is perpendicular to the surface. Note that reception strength is greatest when the receiver and transmitter antennas have the *same* polarization.

The dipole in Fig. 20-1(a) is one-half wavelength, which in free space is given by

$$L = \frac{\lambda}{2}\,(\text{ft}) = \frac{492}{F_{\text{mHz}}} \tag{20-1}$$

Nearer the earth's surface the required length is a little less, on the order of

$$L_{(\text{ft})} = \frac{468}{F_{\text{mHz}}} \tag{20-2}$$

Equation (20-1) holds true when the antenna is physically located many wavelengths from any conductive surface or object, and that includes the earth's surface. Equation (20-2) is an empirically derived approximation that *usually* works for antennas constructed at practical heights above the surface of the earth. Equations appropriate for quarter-wavelength antennas are the same, and are merely Eqs. (20-1) and (20-2) with the velocity factor reduced by one-half, i.e., 246 and 234 instead of 492 and 468, respectively.

Also shown in Fig. 20-1 are the normal voltage and current distributions along the lengths of the radiators. In both points, the *feed point* is a point of low-

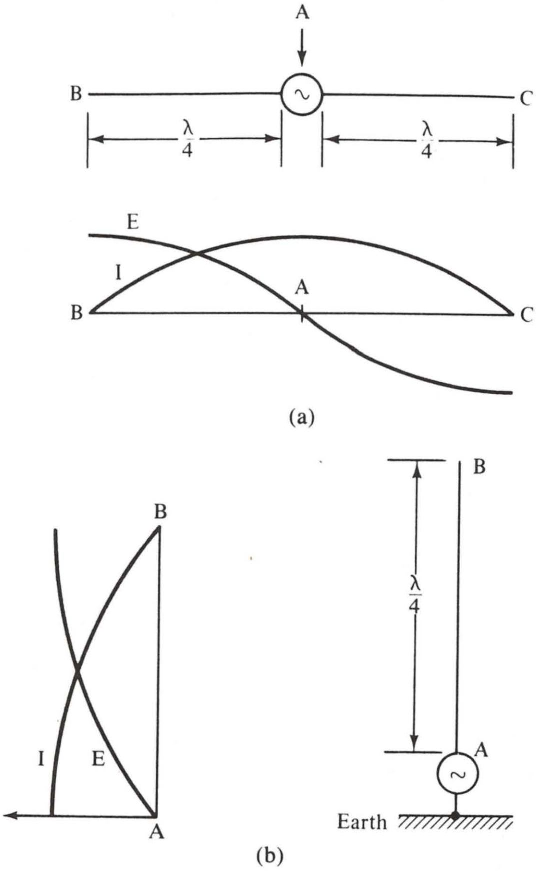


Figure 20-1. (a) Half-wavelength antenna. (b) Quarter-wavelength antenna.

est impedance, i.e., maximum current and minimum voltage. The feed-point impedance is given in the form

$$Z_L = R_0 \pm jX \tag{20-3}$$

where Z_L = the antenna feedpoint impedance in ohms
R_0 = the antenna radiation resistance in ohms
X = the reactive component of impedance
j is the operator $(-1)^{1/2}$

At resonance—that is, when the oscillator frequency is the same as the antenna's natural resonant frequency—the antenna impedance is purely resistive, so $Z_L = R_0$. At frequencies higher than resonance, the phase angle between E and I

becomes *inductive*, so Z_L $R_0 + jX_L$; at frequencies lower than resonance it is capacitive, so $Z_L = R_0 - jX_c$.

20-4 TRANSMISSION LINES

It is rarely practical to locate the oscillator, i.e., radio transmitter, exactly as shown in Fig. 20-1. In most cases, the source will be located at some point that is remote from the antenna feed point, be it just a few feet or a few miles. A *transmission line* is used to transfer power between the source and the load, i.e., from the transmitter output to the antenna feed point (see Fig. 20-2).

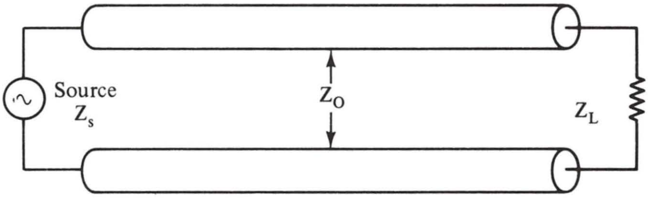


Figure 20-2. Parallel representation of a transmission line.

A transmission line may be a pair of parallel wires, as in Fig. 20-2, or a coaxial cable (which is most often the case in modern radio systems). Note, however, that the true transmission line has properties far more complex than might appear to be the case when only a pair of wires is used. We will leave a complete discussion of transmission lines to a text on antenna systems or an engineering "fields and waves" (i.e., electromagnetics) course. Here we will consider only the *velocity factor* and *characteristic impedance* (also called *surge impedance*) of the line.

The velocity factor arises from the observation that radio waves do not propagate as fast in transmission lines as they do in free space. The velocity of radio signals in space is the speed of light, 3×10^8 m/s, but in a transmission line the velocity will be but a fraction of this speed. This fraction is the velocity factor, and is usually expressed in decimal form (i.e., 0 to 1). Table 20-1 shows the velocity factors for several popular transmission lines.

Table 20-1

Twin-lead: 0.82
Foam-filled coaxial cable: 0.80
Regular dielectric coaxial cable: 0.66

The velocity factor alludes to the fraction of the speed of light at which signals travel in the transmission line. For example, in foam-filled coaxial cable, the signal velocity is 0.80 times the speed of light, or

$$V = (0.80)(3 \times 10^8 \text{ m/s})$$
$$V = 2.4 \times 10^8 \text{ m/s}$$

The velocity factor affects the equations used to calculate electrical half- and quarter-wavelengths. The modified equations for cables are shown below.

Half-wavelength case:

$$L_{(\text{ft})} = \frac{492V}{F_{\text{mHz}}} \tag{20-4a}$$

Quarter-wavelength case:

$$L_{(\text{ft})} = \frac{246V}{F_{\text{mHz}}} \tag{20-4b}$$

where L = the length of the cable in feet (ft)
F = the frequency in megahertz (mHz)
V = the velocity factor (dimensionless)

Example 20-1

Calculate the length in feet of a half-wavelength section of foam-filled coaxial cable at a frequency of 20 mHz.

SOLUTION

$$L = 492V/F \tag{20-4}$$

$$L = \frac{(492)(0.80)}{(20)}$$

$$L = (394)/(20) = \mathbf{19.7\ feet}$$

The characteristic impedance of the transmission line is associated with the distributed inductance and capacitance per unit of length, and is given in the most simple form by

$$Z_0 = \sqrt{L_0/C_0} \tag{20-5}$$

where Z_0 = the characteristic impedance in ohms (Ω)
C_0 = the distributed capacitance per unit length
L_0 = the distributed inductance per unit length

Practical values for Z_0 range from 150 to 1000 ohms for parallel conductor transmission lines, and from 10 to 150 ohms for coaxial conductor transmission lines. Coaxial cable types RG-58/U and RG-59/U have characteristic impedances of approximately 52 ohms and 75 ohms, respectively. Television antenna "twin-lead" has a characteristic impedance of 300 ohms.

Stated another way, the characteristic impedance is equal to the value of the load *resistance* that will permit maximum transfer of power between the source and the load. If any other value of load is used, then not all of the power applied to the line by the source will be absorbed by the load. Some of the power will be reflected back down the transmission line toward the source. Reflected power results in *standing waves* (see Sec. 20-7).

20-5 ANTENNA SYSTEM MEASUREMENTS

The antenna system is potentially very complex, and its measurement is also sometimes complex. Yet, at the same time, it is possible to do much in the way of practical work on antennas using almost naive methods and instruments. This chapter will deal with simple but effective techniques widely used in industry.

Several key parameters are of interest, including *impedance, standing wave ratio, resonance, radiation pattern* (azimuthal and elevation), and *gain.*

20-6 ANTENNA IMPEDANCE

A number of different factors can affect the impedance of a practical antenna, including its physical proximity to the earth's surface and other conductive surfaces. In our simplified world that admits only half-wavelength horizontal dipoles and quarter-wavelength vertical monopoles, we find that "ideal" impedance values are 52 ohms for the vertical and 73 ohms for the horizontal antenna. But in actual practice, real-world antennas have feed-point impedances that vary considerably from these ideals. The horizontal dipole, for example, will exhibit a radiation resistance between 5 ohms and 120 ohms for heights above ground of 1/25 to 1 wavelength above ground. The 73-ohms figure is merely a theoretical impedance in free space. Similarly, the quarter-wavelength vertical radiator has a nominal or theoretical impedance of 52 ohms, but exhibits impedances between 5 and 80 ohms (mobile installations, in which such antennas are particularly popular, usually show feed-point impedances of 10–30 ohms). These values of impedance represent only the radiation resistance component of the overall impedance, which is given as

$$R_0 = \frac{E_0}{I_0} \tag{20-6}$$

where R_0 = the radiation resistance in ohms
E_0 = the feedpoint potential in volts
I_0 = the feedpoint current in amperes

The reactive component is zero in ideal, resonant antennas, but usually has a value between 0 and ±100 ohms in real antennas.

20-7 "STANDING WAVES" AND SWR

Standing waves on antennas and transmission lines are simple phenomena, but are often misunderstood. A certain mystique has built up, especially among Citizens Band operators, over the matter of *standing wave ratio* (SWR) and what it means in terms of radio system performance. Many people spend a great deal of money and time attempting to idealize the SWR on an antenna without realizing that little improvement results in their ability to communicate with other stations.

Before discussing SWR, however, let us first determine the nature of standing waves. We may do this best by mechanical analogy, using a *model* that is easy to understand and interpret. Incidentally, modeling, although it is a valid teaching tool, is often poorly done. But in the case of standing waves, the model is another standing wave, and the behavior of the two types are nearly identical; the experiment using the model can be done by almost anybody who has a rope and a support to attach it to.

Our rope model is shown in Fig. 20-3, and consists of a rope anchored at one end to a wall or other immovable, permanent structure. The rope is held *taut* by a person [Fig. 20-3(a)] at the other end.

A *wave* is imparted to the rope [Fig. 20-3(b)] by a quick up and down motion of the wrist. This wave travels at a given velocity towards the wall. When the wave hits the wall end of the rope, it is *reflected* back toward the source end of the rope. Note that a phase reversal took place when the wave was reflected.

If the rope were ideal, and both the wall and the operator's hand were perfectly rigid, then the wave would reflect back and forth along the line undiminished as time went on. But, like electrical circuits, there are losses that absorb energy and reduce the amplitude of the wave.

The situation in Fig. 20-3 is analogous to applying a *pulse* to an electrical transmission line. But what happens when a *continuous* oscillation is applied to

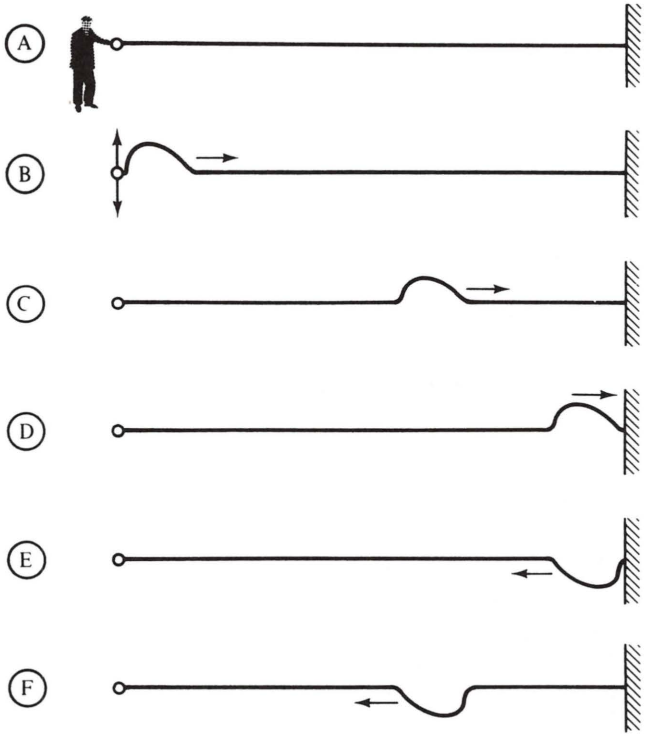


Figure 20-3. Concept of a reflected wave using a rope analogy.

the line? Consider Fig. 20-4. The source end of the rope is moved up and down in a precise, well-timed manner, thereby imparting sine waves to the rope. After the first wave has reflected from the wall end of the rope, we will actually have two waves: a *forward* wave traveling from left to right and a *reflected* wave that travels right to left. The wave amplitude at any given point along the rope will be the algebraic sum of these two waves. When viewed from the side, the rope will take on a blurred appearance, showing the nodes and antinodes of Fig. 20-4. These are known as *standing waves* because the nodes and antinodes are stationary; they do not move relative to the length of the rope.

Figure 20-5 shows three situations where there are standing waves on a radio transmission line (such as Fig. 20-2). Figure 20-5(a) shows the situation when the forward wave is *totally* reflected back toward the source; the case when the transmission line is *shorted* ($Z_L = 0$) or totally *open* ($Z_L = \infty$). Interestingly enough, the pattern for both conditions is the same, except for the location of the nodes and antinodes. If the line is shorted, then the voltage at the load end is zero, so the nodes, i.e., points of minimum voltage, are spaced at half-

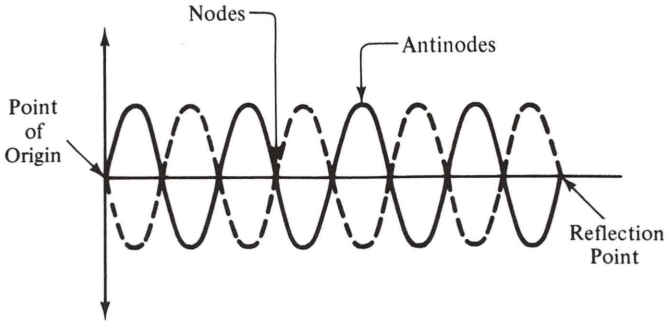


Figure 20-4. Standing waves on a line.

wavelength intervals along the line back from the short. But if the line is open, the voltage at the load end is maximum, so the antinodes, i.e., points of voltage maxima, are spaced at half-wavelength intervals back from the load end of the cable. In other words, the two patterns are identical, except for a 90-degree shift in the relative positions of the nodes and antinodes. The particular pattern selected for Fig. 20-5(a) is representative of the shorted condition.

Most antenna systems, however, do not totally reflect the forward wave; some of it is radiated into space. This means that the reflected wave has a lower amplitude than the forward wave, so the cancellation is not total. The graph in Fig. 20-5(b) shows this situation; note that the voltage at the nodes is *not* zero. The *standing wave ratio* (SWR) is given by

$$\text{SWR} = \frac{E_{\max}}{E_{\min}} \tag{20-7}$$

or, if the current is measured instead of voltage,

$$\text{SWR} = \frac{I_{\max}}{I_{\min}} \tag{20-8}$$

Some people designate these measurements as *VSWR* and *ISWR,* respectively, but the numbers are the same. We may also calculate SWR from

$$\text{SWR} = \frac{Z_0}{Z_L} \quad (Z_0 > Z_L) \tag{20-9}$$

$$\text{SWR} = \frac{Z_L}{Z_0} \quad (Z_0 < Z_L) \tag{20-10}$$

$$\mathrm{SWR} = \frac{1 + \sqrt{P_r/P_f}}{1 - \sqrt{P_r/P_f}} \tag{20-11}$$

and

$$\mathrm{SWR} = \frac{E_f + E_r}{E_f - E_r} \tag{20-12}$$

where Z_0 = the characteristic impedance of the transmission line in ohms
Z_L = the impedance of the load
P_r = the power reflected from the load
P_f = the forward power applied to the transmission line by the source
E_r = the voltage component of the reflected wave
E_f = the voltage component of the forward wave

The last case is shown in Fig. 20-5(c), and represents the situation where $Z_L = Z_0$. In this condition, the forward power is totally absorbed by the load, and none is reflected. In this case, there are no nodes or antinodes, and the line is said to be *flat*. Ideally, this is what is desired in a perfect antenna-transmission line system, and represents the case where the SWR is 1:1.

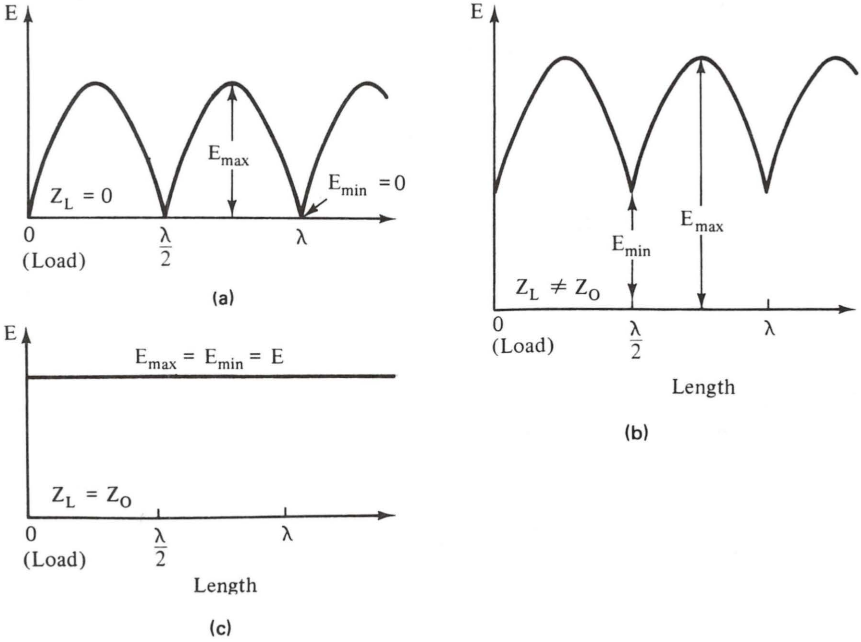


Figure 20-5. (a) Standing waves for $Z_L = 0$. (b) Standing waves for $Z_L \neq Z_o$. (c) Standing waves for $Z_L = Z_o$.

20-8 MEASURING SWR

Any of the equations for SWR presented so far can be used as the basis for an SWR measurement system. We could, for example, measure the impedance of the antenna, and then use Equations (20-9) or (20-10). Or we could actually measure the power, voltage, or currents in the forward and reflected waves.

All of the methods work, and one can find instruments that are based on just about any of them. In most low-cost VSWR indicators, however, either voltage or current is used, and that produces a problem that is the root of one of the most persistent myths about SWR adjustment and measurement. It is the *erroneous* belief of many people that SWR can be "adjusted" by trimming the transmission line length. This is a pernicious and vicious idea that ought to be consigned forthwith to the city dump. This idea is not true, but the belief that it is true persists because the nature of voltage- and current-based SWR meters make it *appear* true to the unknowing observer. When this type of instrument is used the *numbers* obtained are valid *only* when SWR is measured at the antenna feed point, or at half-wavelength intervals along the line back from the load. In other words, for the numbers to be valid, the transmission line bewtween the SWR meter and the antenna must have a length L of

$$L = \frac{492NV}{F} \tag{20-13}$$

where N is a convenient integer (i.e., 1, 2, 3, . . .) and all other terms are as defined in Eq. (20-4a).

In addition to the problem just discussed, simple SWR meters also suffer greatly in accuracy when used in systems where the antenna is incorrectly built, and has a grossly improper length. Such antennas are highly reactive, and so they will interfere with the correct measurement of SWR. Attempting to obtain accurate readings where the SWR is very high, if low-cost instruments based on voltage or current are used, is a lot like trying to skin an amoeba.

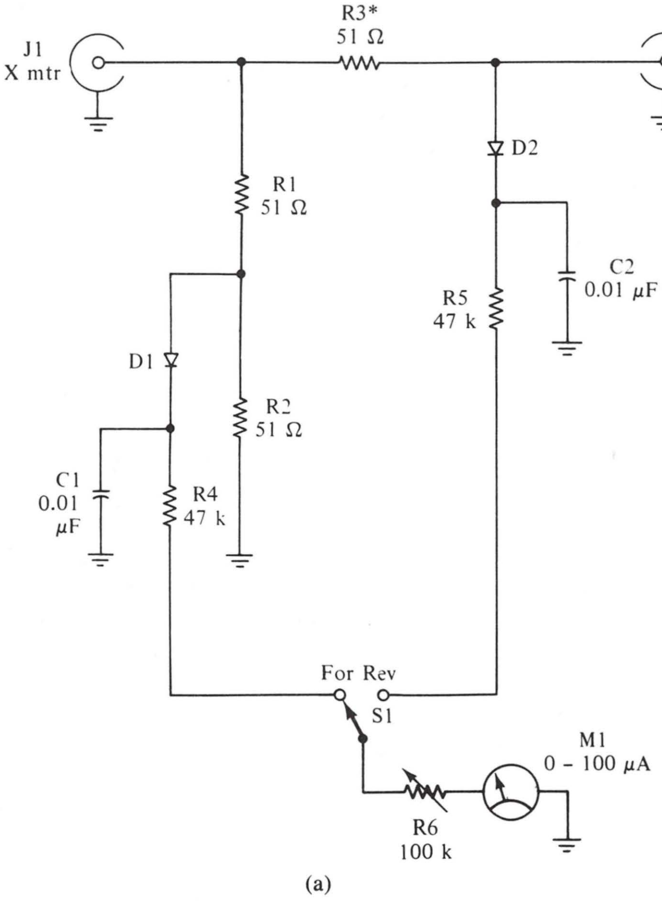
The low-cost SWR meter is a valid instrument if its limitations are understood. If the SWR is not too high, and if the user understands that it is a *relative* indicator, then it will prove very useful in correctly adjusting radio antenna systems. The correct use is to adjust the antenna length, or the L and C components in an antenna-matching network at its base, for a *minimum* SWR reading. The minimum will always be located at the same adjustment of the antenna components, even though we know not to take the numerical indication too seriously unless the line length is correct and the antenna is not terribly reactive.

If a true RF wattmeter is used to measure forward and reflected power, so that Eq. (20-11) may be applied, it is found that SWR is totally independent of line length. Some RF wattmeter manufacturers publish an SWR nomograph based on Eq. (20-11) to make this calculation easier (Fig. 20-7, to be discussed).

20-9 SWR METERS

In addition to the RF power meters discussed in Chapter 22, there are several low-cost meters usually employed to measure SWR. Examples of some circuits are shown in Fig. 20-6. These instruments are voltage- and current-based, respectively.

The bridge SWR meter is shown in Fig. 20-6(a). It is a resistance Wheatstone bridge of the type discussed in Chapter 3. The bridge arms consist of resistors $R1$ through $R3$, and the radiation resistance of the antenna. The transmitter, or a low-power RF source such as a signal generator, produces the excitation potential. Note that this is a *resistance* bridge, so it will not measure the reactive component.



(a)

Figure 20-6. (a) Simple bridge-type VSWR meter.

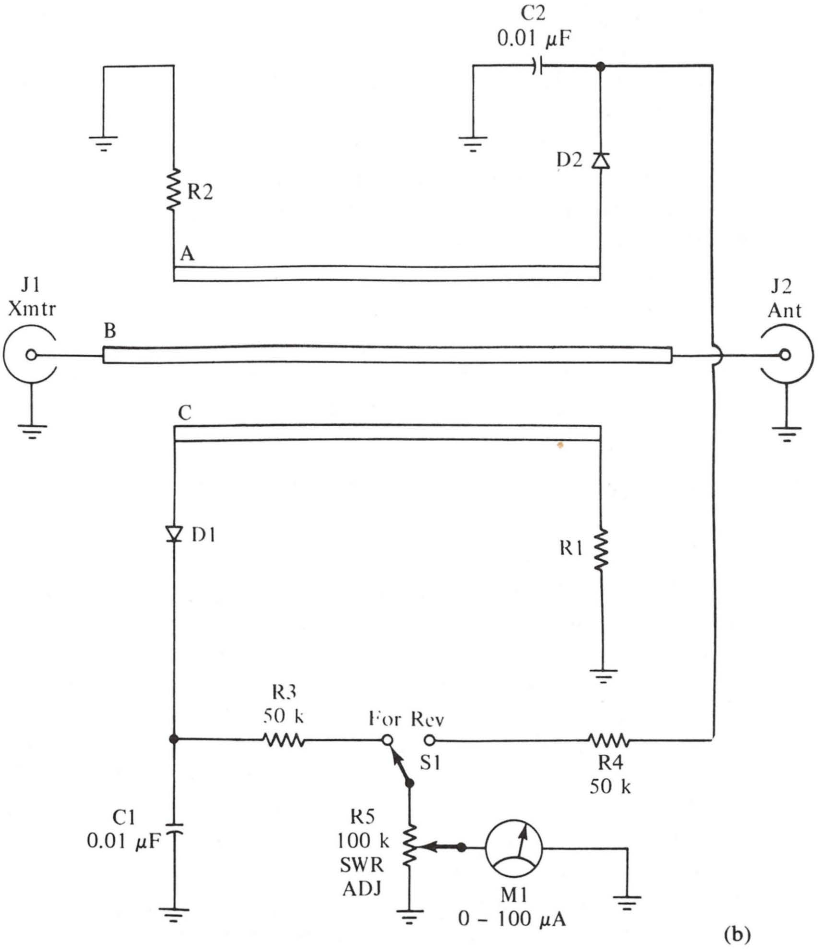


Figure 20-6. ***(cont.)*** (b) "Monimatch"-type VSWR meter.

The circuit shown in Fig. 20-6(a) is set up to measure the radiation resistance of 50-ohm antenna systems because $R3$ is set to 51 ohms. If this resistor is changed to 75 ohms, then the bridge will also work accurately in 75-ohm systems. The operation of the bridge is as follows.

1. Set $S1$ to *forward*.
2. Adjust $R6$ for a full-scale deflection.
3. Set $S1$ to reflected, and read SWR from special meter scale.

Another form of SWR meter is shown in Fig. 20-6(b), and has the advantage over the previous type that it may be left in the line while transmitting. The heart of this instrument is the pick-up unit, consisting of the three conductors labeled A through C.

Conductor B is the center conductor of the transmission line, while A and C are identical to each other and spaced equally distant from B. In older units conductor B was a piece of coaxial cable, while conductors A and C were small-gage enameled wires threaded between the braid and inner insulation of the coaxial cable. In most modern instruments, however, the pick-up unit is made from foil patterns on a printed circuit board.

Resistors $R1$ and $R2$ are selected to have a resistance equal to the characteristic impedance of the transmission line, i.e., 50 or 75 ohms. In some units a trade-off value of 68 ohms is used so that the instrument may be used with either type of transmission line with but minimal error. Except for these resistors, the detector circuits and their operation are the same as on the bridge type of meter.

20-10 USING POWER METERS TO MEASURE SWR

SWR can also be measured with an RF wattmeter that is equipped with a direction coupler device so that forward and reflected power can be measured. We can then use the power levels obtained on the wattmeter in Eq. (20-11) to determine SWR. Bird Electronics, a manufacturer of professional RF wattmeters, publishes the nomograph shown in Fig. 20-7, and this chart is based on Eq. (20-11). Two different ranges are shown in Figs. 20-7(a) and 20-7(b), respectively.

It is also possible to use the radiation resistance of the antenna, as measured on an antenna impedance bridge. Once this value is determined, then either Eq. (20-9) or (20-10) may be used, as appropriate, to compute the SWR.

There are also some computing SWR meters on the market. Most of these use a pick-off unit such as Fig. 20-6(b), or a related mechanism in which a toroidal (i.e., doughnut-shaped) core current transformer replaces the pick-up unit, but also includes an analog circuit that performs the three-step procedure given earlier automatically.

20-11 ANTENNA IMPEDANCE BRIDGES

There are several types of antenna impedance bridge on the market, but most are based on the Wheatstone bridge circuit, as in Fig. 20-8(a). Typically, impedance $Z4$ is the antenna impedance, $Z3$ is fixed, and both $Z1$ and $Z2$ are variable.

A simple resistance bridge of the form shown previously in Fig. 20-6(a) will often be used, but this type of circuit measures only the radiation resistance component of impedance, and ignores altogether the reactive portion. But this is not necessarily disadvantageous, because the reactive component tends to disappear in simple antennas that are made resonant; a procedure for resonating the antenna can be performed without a complex impedance bridge.

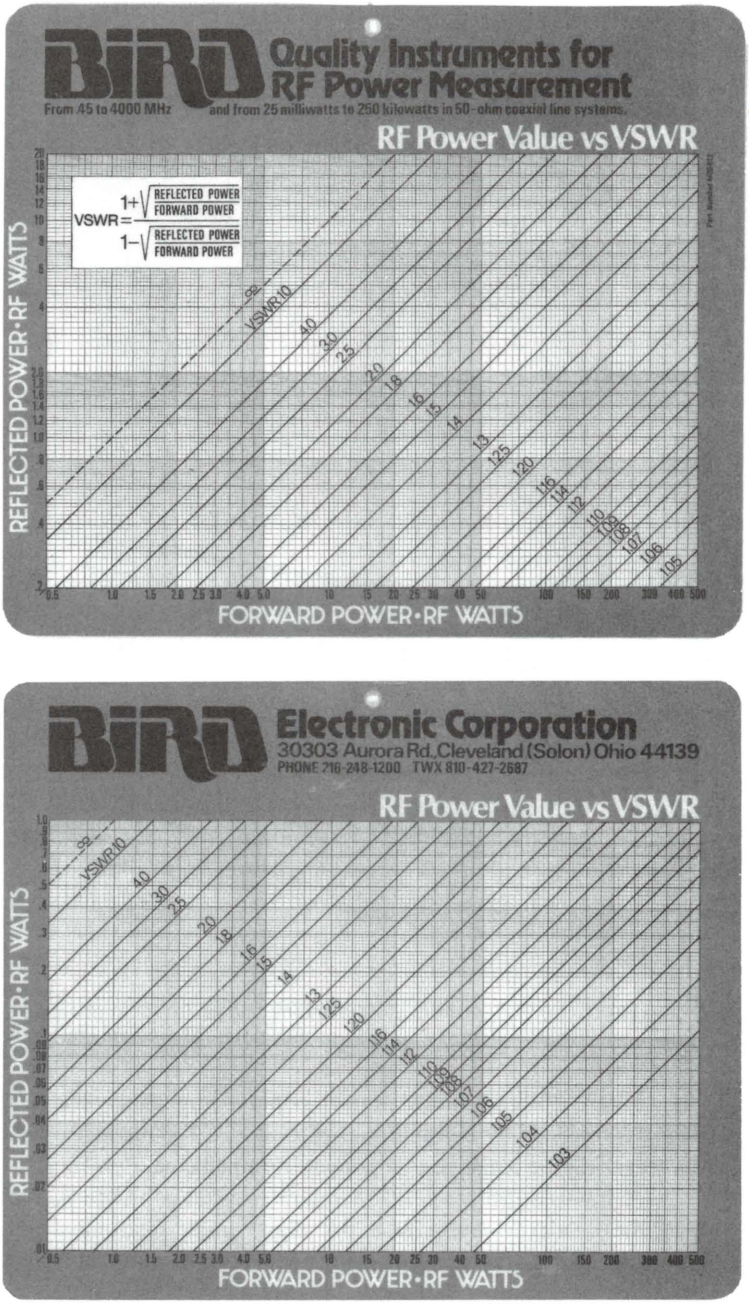


Figure 20-7. Nomograph for computing VSWR from RF power. (Courtesy of Bird Electronics Corp.)

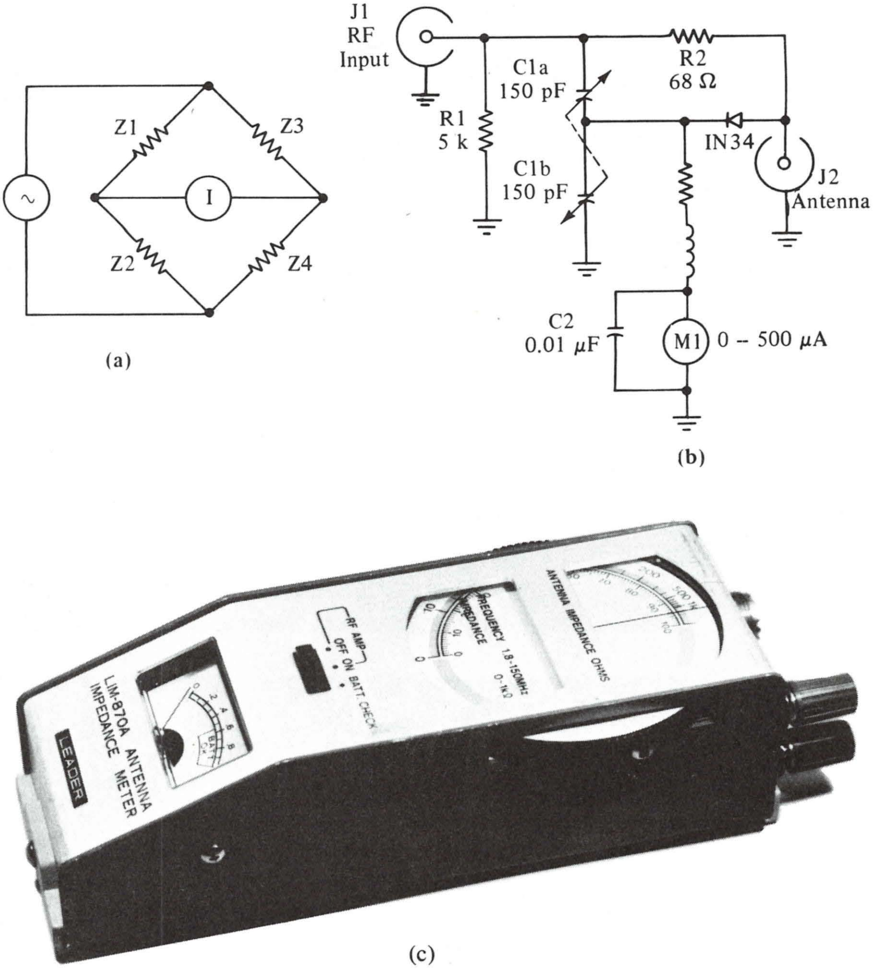
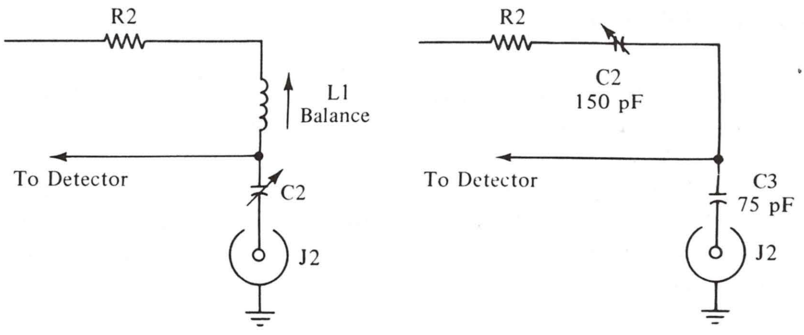


Figure 20-8. (a) General form of Wheatstone bridge. (b) RF impedance bridge for antenna measurements. (c) Low-cost commercial antenna impedance bridge.

Figure 20-8(b) shows another popular impedance bridge circuit in which $Z1/Z2$ is replaced by a differential capacitor, $C1$. This instrument is also limited to measuring the resistive component of impedance.

A low-cost commercial radiation resistance bridge is shown in Fig. 20-8(c), the Leader Model LIM-870A. This simple instrument will determine the value of radiation resistances between 0 and 1000 ohms, at frequencies up to 150 mHz.

None of the bridges shown so far will measure the antenna's reactive component of impedance; only the antenna's radiation resistance can be actually



(e) (d)

Figure 20-8. *(cont.)* (d) and (e) Modifications of resistance bridge to allow measurement of complex impedances: (d) inductive, (e) capacitive.

measured numerically. But an experienced operator can usually *spot* highly reactive conditions by observing the depth and sharpness of the null. A resistive antenna shows a deep, narrow null, whereas a reactive antenna exhibits a shallower, more broad null. This type of determination is, at best, subjective and requires skill in interpretation. But it can aid by showing that the antenna is not resonant.

If it is necessary to measure the reactive component of antenna impedance, then a modified bridge capable of measuring ac impedances (Chapter 3) is required. In some cases, a balancing inductor and calibrated variable capacitor are placed in series with the antenna impedance, forming a series-resonant circuit. This combination [see Fig. 20-8(e)] is adjusted so that $L1$ (i.e., the *balance* control) resonates with $C2$ at the excitation frequency when $C2$ is exactly in the middle of its range. The impedance of $L1/C2$ under this condition is entirely resistive, and will have a very low value.

Another technique that is often employed is shown in Fig. 20-8(d), and consists of two capacitors, $C2$ and $C3$. Capacitor $C2$ is a calibrated variable model, and must be of the "straight line capacitance" type of construction. Capacitor $C3$ is a precision fixed capacitor, and has a value that is equal to $\frac{1}{2}C2$. When $C2$ is set to midrange, then, these capacitances cancel, i.e., are balanced in the bridge circuit. The dial connected to $C2$ will read zero at midscale. If the antenna has a reactive component, then $C2$ must be offset higher (X_L) or lower (X_c), depending upon whether the reactance is inductive or capacitive.

Note that these measurements are also subject to the same constraints as SWR measurements: The bridge must be connected into the system at the antenna feedpoint, or through a piece of transmission line that is a half-wavelength long.

20-12 NOISE BRIDGES

The impedance noise bridge, low-cost commercial example shown in Fig. 20-9, is a low-cost method for testing antenna systems.

A representative circuit is shown in Fig. 20-10. The noise source is a reverse-biased zener diode, and produces a wide-spectrum pseudogaussian noise signal that is amplified in a wideband (i.e., -3 dB at over 150 mHz) amplifier. The noise output from the amplifier is applied to the primary of a special transformer, $T1$.

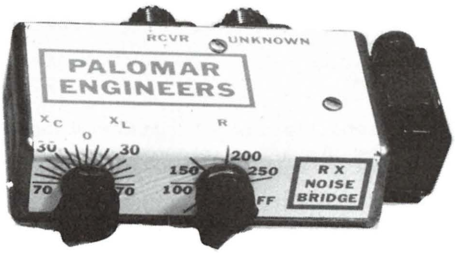


Figure 20-9. Low-cost R-X noise bridge.

This transformer ($T1$) has a 1:1:1 ratio, and windings $L1$ through $L3$ are *trifilar wound* on a toroid core (see inset to Fig. 20-10). The dots in Fig. 20-10 indicate the same ends of $L1$, $L2$, and $L3$.

One end of $L2$ is connected to a noninductive potentiometer that has a dial calibrated in ohms, from zero to full scale (in this example, 250 ohms), and a variable capacitor connected to a dial marked in noncalibrated units of reactance from X_c-0-X_L. As in Fig. 20-8(d), $C1$ is balanced by $C2$ so that at $X = 0$, $C1 = C2$.

The detector used with a noise bridge is a general coverage communications receiver, preferably a model equipped with an "S-meter," or at least an AGC voltage meter. The receiver will show a high noise level at all frequencies *except* the resonant frequency of the antenna.

The noise bridge can be used to find the resonant frequency of an antenna, the correct physical length for a piece of transmission line that must be cut to an electrical half-wavelength, and certain other applications.

20-13 FINDING AN ANTENNA'S RESONANT FREQUENCY

There are two ways to use a noise bridge to find the resonant frequency of an antenna. For those bridges that do not use $C1$ and $C2$, the procedure is as follows.

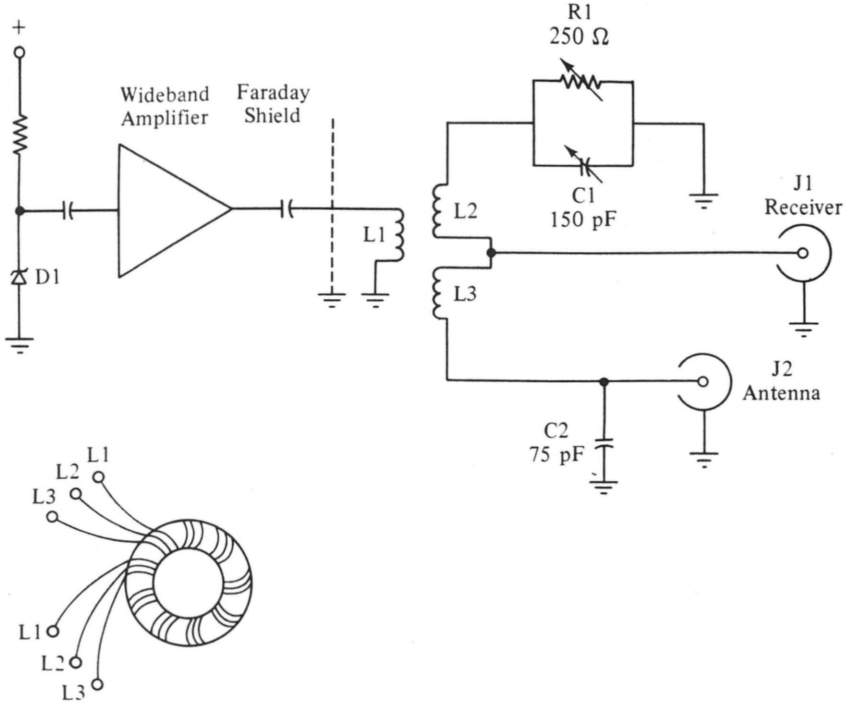


Figure 20-10. Circuit for noise bridge.

1. Set $R1$ to the anticipated radiation resistance of the antenna; tune the receiver to the anticipated resonant frequency.
2. Adjust the receiver tuning *very slowly* above and below the antenna's anticipated resonant point until a null is heard in the noise level, or a dip is noted on the receiver's S-meter.
3. Adjust the receiver dial and $R1$ for the deepest null in the noise level, or minimum S-meter reading. These controls are slightly interactive. The receiver dial now indicates the resonant frequency, while the resistance of $R1$ is equal to the antenna's radiation resistance.

On models such as Fig. 20-10, which have $C1$ as a reactance control, use the procedure below.

1. Set $R1$ to the anticipated radiation resistance, set $C1$ to $X = 0$ (i.e., midscale), and tune the receiver to the anticipated resonant frequency.
2. Adjust $C1$ for a null, or minimum reading on the S-meter.

3. If the null occurs when $C1$ is in the X_c region, the antenna is too long. But if the null is in the X_L region, then it is too short.
   a. $X = X_c$: tune the receiver *downband* to locate f.
   b. $X = X_L$: tune the receiver upband to locate f.
4. Adjust $C1$, $R1$, and the receiver dial for best null, as in the previous example.

20-14 FIND λ/2 LENGTH OF COAXIAL CABLE

We learned from Eq. (20-13) that a piece of transmission line will vary from a physical half-wavelength because of the velocity factor V. This phenomenon is due to the slower velocity of the radio signal in coaxial cable as compared with free space. The correct physical length to make an electrical half-wavelength can be found by using a noise bridge and Eq. (20-13) in the procedure below.

1. Use Eq. (20-13) to calculate the required length, and then cut a length of cable that is slightly *longer*.
2. Set X to 0, and $R1$ (in Fig. 20-10) to some resistance slightly above zero.
3. Connect one end of the cable to the noise bridge antenna port, and short the other end.
4. Tune the receiver to the frequency at which the cable must be a half-wavelength, and then tune downband until a pronounced null is heard.
5. Patiently and slowly, trim the cable length, and then reshort the end after each cut, until the null moves upband to the desired frequency.
6. The result may be checked by removing the short and terminating the cable with a carbon or other noninductive resistance in the 10-to-250-ohms range. Adjust $R1$ for best null at the desired frequency. The dial on $R1$ should indicate a resistance that is very nearly the value of the resistor used to terminate the cable.

20-15 USE OF THE NOISE BRIDGE TO FIND VELOCITY FACTOR OF COAXIAL CABLE

We may rearrange Eq. (20-13) to give us the velocity factor V in terms of line length and frequency. Namely,

$$V = \frac{LF}{492N} \tag{20-14}$$

where V = the velocity factor and has a value between 0 and 1
L = the length in feet (ft)
F = the frequency in megahertz (mHz)
N = any integer (for convenience, set $N = 1$)

Procedure

1. Cut a sample length *(L)* of the coaxial cable being tested. It should have a length that is an electrical half-wavelength at some frequency within the tuning range of the receiver. Measure the length in feet.
2. Calculate the free-space wavelength of the cable, using the formula $F = 492/L$. Tune the receiver to this frequency.
3. Connect one end of the cable to the noise bridge, and *short* the other end.
4. Set $R1$ on the noise bridge to a value that is slightly higher than zero.
5. Tune the receiver downband until a null is obtained.
6. Using the length L found in Step 1, and the frequency found in Step 5, use Eq. (20-14) to calculate V.

20-16 SWEEP TECHNIQUES

A frequency-modulated (FM) or sawtooth-modulated sweep generator can also be used to test transmission lines and antennas for resonant frequency. Figure 20-11(a) shows the equipment configuration to make this type of measurement. The signals from the sweep generator and a crystal-controlled marker generator (used to identify points on the frequency display) are combined in a resistor-diode adder circuit. The output of the adder is a dc voltage, and remains relatively constant at frequencies removed from resonance. But at frequencies near resonance, a null or dip is noted. A sample oscilloscope trace for a piece of coaxial cable shorted at the half-wavelength point is shown in Fig. 20-11(b). The marker "pip" in the null indicates that the cable is indeed cut to the correct frequency.

20-17 POLAR PATTERNS

An antenna may or may not be omnidirectional, depending upon its physical design. In fact, antenna *gain* is nothing more than directing the power in one or two directions instead of in all directions.

Polar patterns are often created in an anchroic test chamber in which the antenna, or a VHF model (i.e., scaled down in size), is surrounded by cones of RF absorptive material that reduces reflections (see Fig. 20-12). Alternatively,

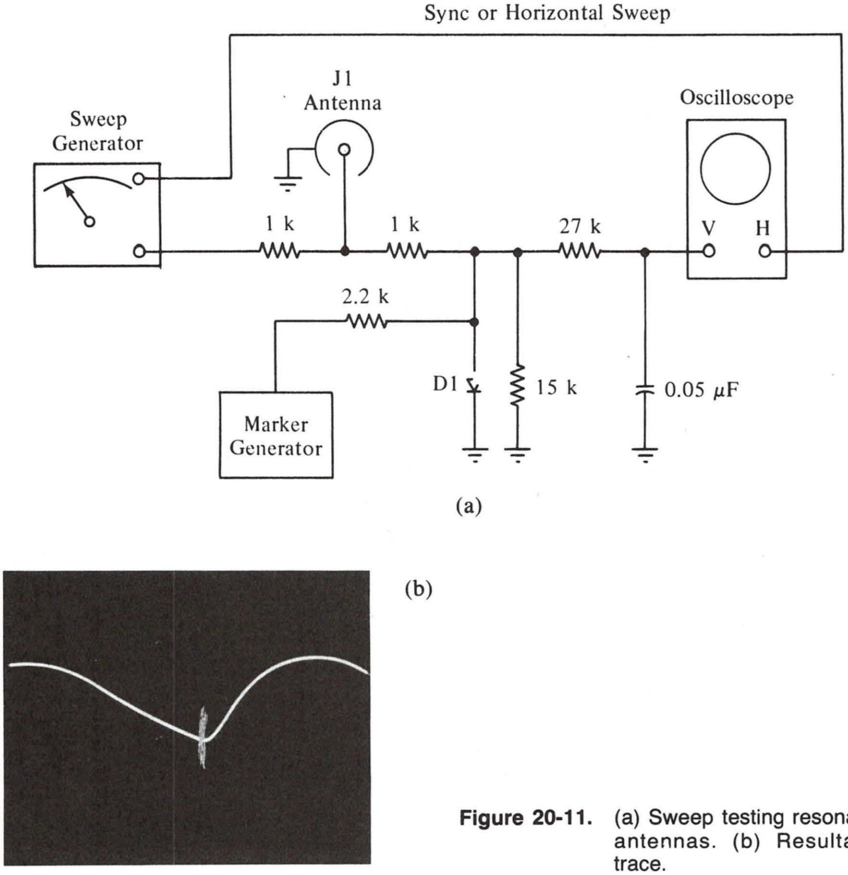


Figure 20-11. (a) Sweep testing resonant antennas. (b) Resultant trace.

the antenna is in an outdoor antenna range where there are no objects for many wavelengths.

Figure 20-12 shows a model of an automobile with a pair of quarter-wavelength vertical monopoles mounted one-eighth wavelength apart on the trunk. This model is a 300-mHz scale-down of a 30-mHz (i.e., CB) cophased antenna system.

The method used to make the polar plot is diagrammed in Fig. 20-13(a). The antenna, or the model containing the antenna, is mounted on a motor-driven platform. The platform position transducer commands a servomotor that causes a turntable chart recorder to rotate. Polar coordinate graph paper is mounted on the turntable, and a radio receiver tuned to the test frequency drives the pen position amplifier. A *distant* (i.e., many wavelengths) RF emitter, i.e., another antenna that is connected to a transmitter, sends out a signal whose strength is plotted as a function of radius on the graph paper. The servo system causes the paper to rotate as the model rotates, thereby creating the polar radiation plot of that antenna. Figure 20-13(b) shows typical antenna patterns.

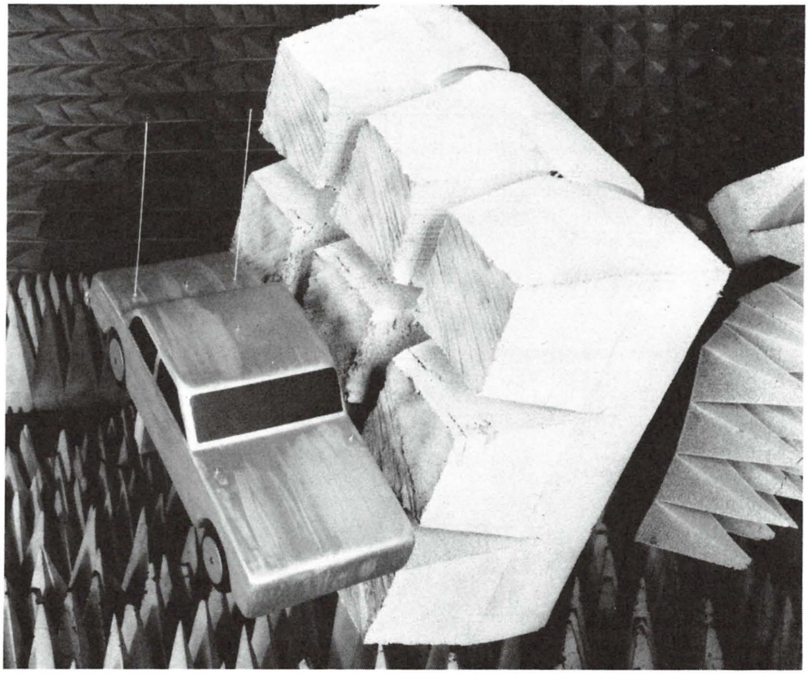


Figure 20-12. Antenna model in reflection-free test chamber. (NASA photo—courtesy NASA.)

20-18 ANTENNA GAIN

The subject of antenna gain measurement usually provokes no small amount of controversy. The problem revolves around methodology and definitions, and these vary considerably from one manufacturer to another, so very different results are reported for the same general types of radio antenna. Some of this difference may be due to honest differences of opinion within the antenna engineering profession, but in many cases it appears to be the result of a little of the aforementioned "creative specification writing" by the company's advertising department. Higher gain for less money, after all, enhances customer appeal so much that some CB antenna companies now claim ridiculous forward-gain figures. We will endeavor to offer some sensible techniques that allow relative comparisons between different types of antennas.

Two different types of figures are often considered important in the comparison of gain antennas: *forward gain,* and either *front-to-back* or *front-to-side* ratios (which depends upon antenna design).

The antenna gain is achieved by redirecting the energy from the transmitter in one direction, instead of in all directions (in the vertical monopole), or in two directions (in the horizontal dipole). The total energy in the field remains the

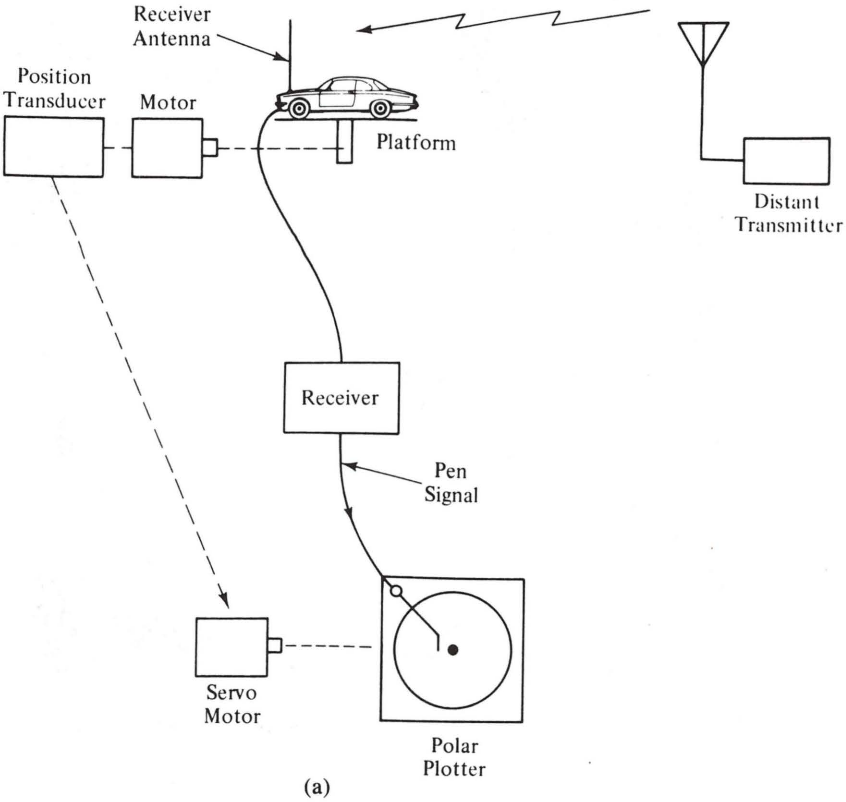
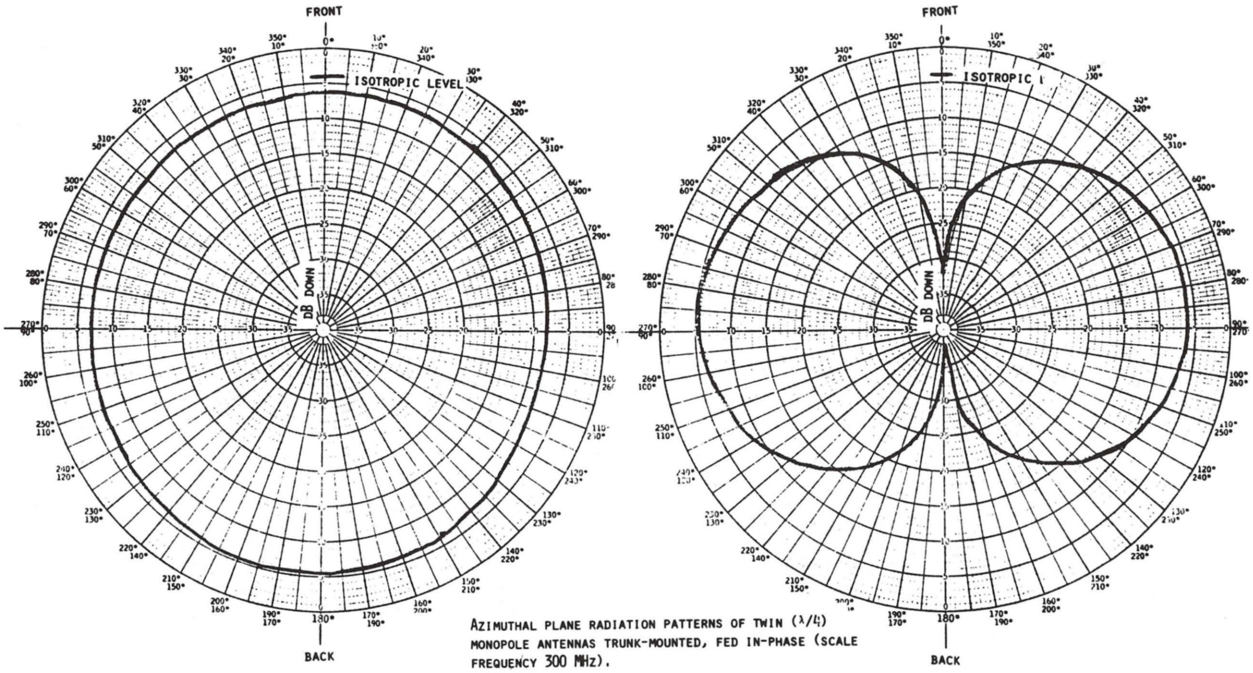


Figure 20-13. (a) Antenna range for plotting polar patterns.

same, but it is redistributed to increase the field intensity in the favored direction(s). The polar pattern shows us the favored directions, and the same pattern holds true for both transmit and receive conditions. In the receive condition, we note that the antenna pattern causes the receiver to be *more sensitive* in some directions than in others. In addition to increasing the signal arriving from the forward direction, it will also reject interfering signals arriving from the direction of the nulls in the polar pattern. This capability can be judged on a receiver. A single half-wavelength dipole is bidirectional (that is, it has a figure-eight pattern), while a beam or yagi is unidirectional. We would, therefore, measure also the front-to-side ratio on a dipole and front-to-back ratio on a beam.

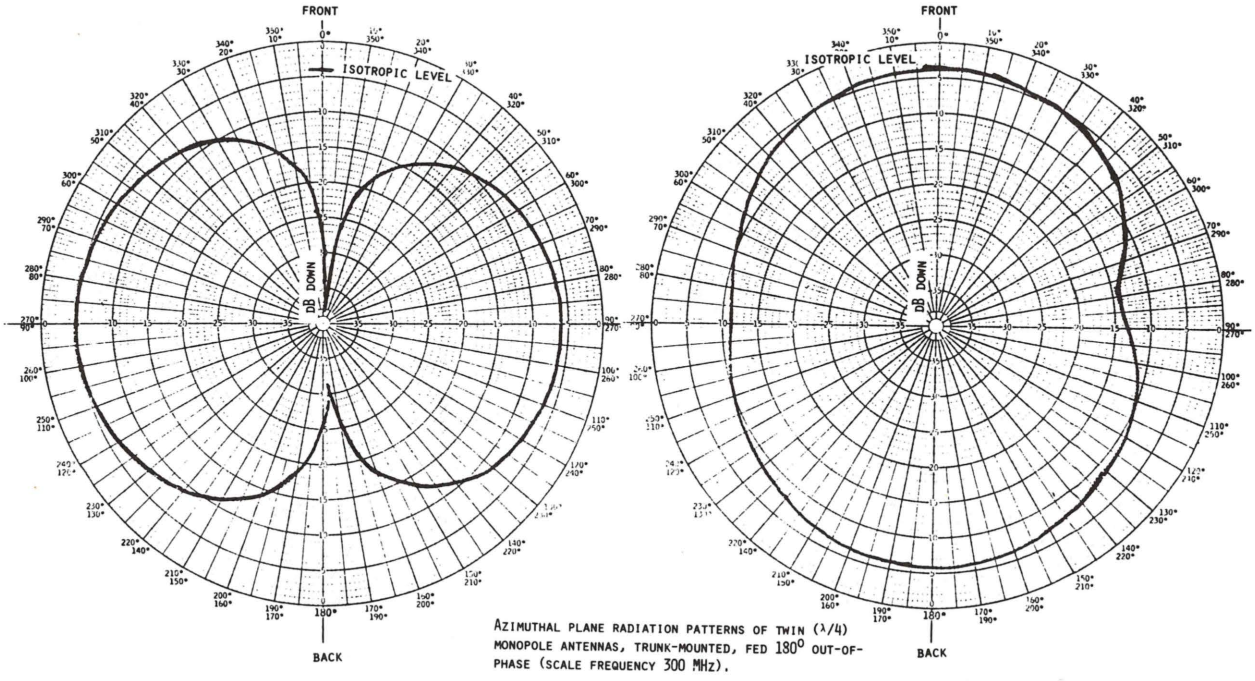
Figure 20-14(a) shows a test configuration for roughly measuring these ratios. The antenna being tested is installed in a way that allows it to be rotated. A constant output signal source is located downrange many wavelengths away. The signal strengths at the receiver are measured for two conditions: main lobe



AZIMUTHAL PLANE RADIATION PATTERNS OF TWIN (λ/4) MONOPOLE ANTENNAS TRUNK-MOUNTED, FED IN-PHASE (SCALE FREQUENCY 300 MHz).

(b)

Figure 20-13. *(cont.)* (b) Polar patterns for several antennas. (NASA photo—courtesy NASA.)



AZIMUTHAL PLANE RADIATION PATTERNS OF TWIN ($\lambda/4$) MONOPOLE ANTENNAS, TRUNK-MOUNTED, FED 180° OUT-OF-PHASE (SCALE FREQUENCY 300 MHz).

(c)

Figure 20-13. *(cont.)* (b) Polar patterns for several antennas. (NASA photo—courtesy NASA.)

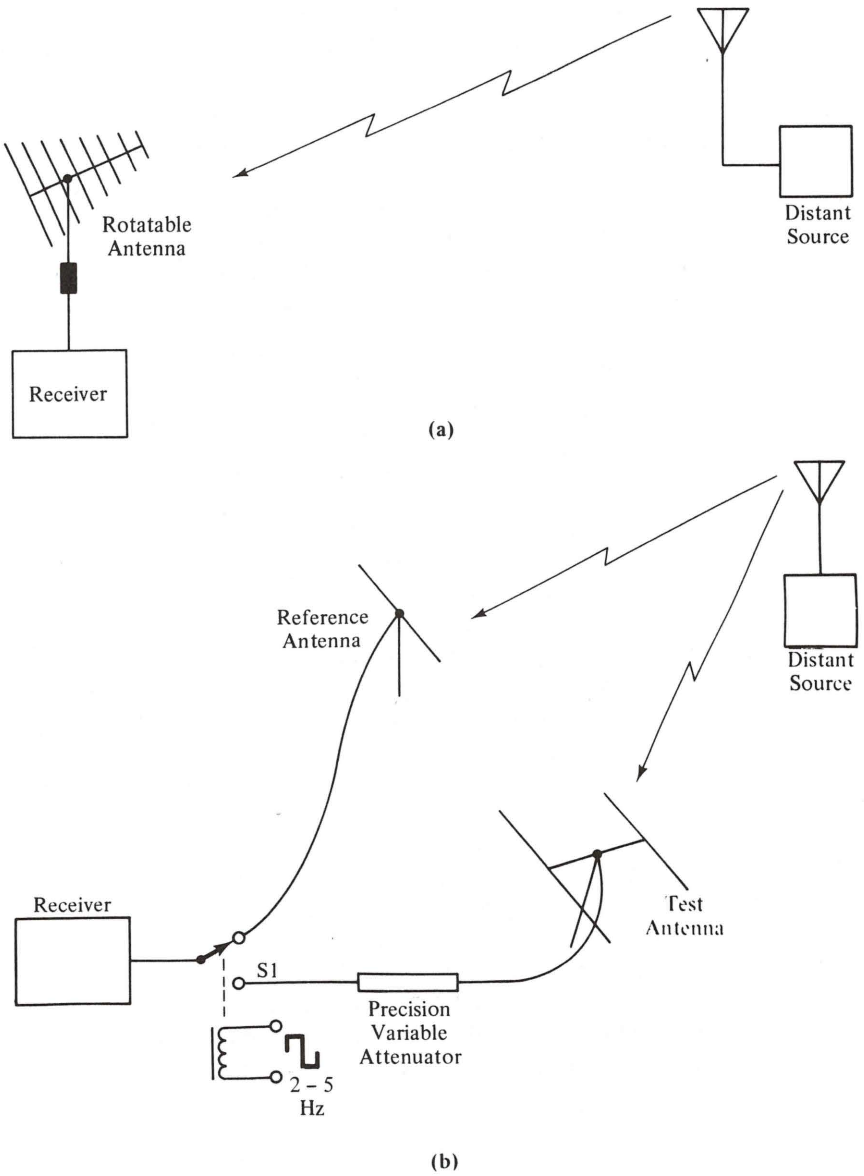


Figure 20-14. (a) Testing for polar pattern at fixed location. (b) Measuring comparative antenna gain.

looking into the signal source, and the null looking into the signal source. The levels obtained are compared, and the difference in decibels is noted.

The receiver's S-meter will give a rough indication in decibels, provided that its S-unit/dB calibration factor is known. A somewhat superior technique, however, is to use a precision RF attenuator that has calibrated, selectable steps. The attenuator is used in conjunction with a receiver S-meter, but the S-meter is used merely as a relative indicator. First note the precise S-meter reading when the antenna null is pointed toward the source. Next, insert the step attenuator into the line and point the antenna main lobe towards the signal source (i.e., note the point of maximum signal strength as the antenna is rotated). The signal level reported by the S-meter should have increased considerably. Now, adjust the step attenuator until the same S-meter is obtained as was found when the null was pointed at the antenna. The ratio is given by the amount of attenuation, in dB, that was required to bring the two signal strengths into equality.

Either of these methods, S-meter or attenuator reading, results in only an approximate result, so they should be repeated many times over the course of several days, and then averaged. The principal problem is in keeping the intensity of the field from the RF emitter constant over the entire measurement period. This problem becomes especially acute when a distant broadcasting station is used as the source because of ionospheric problems and fading.

Antenna gain is measured by using a test setup similar to Fig. 20-14(b). Here we are referencing a test antenna to a standard antenna, such as a dipole, that is oriented in the same direction. Again, we use a distant RF emitter as the signal source, often a broadcasting station signal.

Switch $S1$ is an electrically driven coaxial relay that is pulsed to alternately connect both antennas to the receiver at a 2–5-Hz rate. A precision attenuator (as used above) is inserted in series with the transmission line of the higher gain antenna.

Since the receiver input is alternating back and forth between two antennas with different gains, the S-meter will oscillate at the switching rate, attempting to measure the signal strength from first the reference antenna and then the test antenna. The degree of attenuation required to cause the S-meter to cease oscillating indicates the relative difference in dB between the two antennas. For example, if a three-element beam were compared with a dipole cut to the same frequency and oriented in the same direction, and 4.6 dB of attenuation in the feed line from the beam were required to create equal signal levels, then the "gain" of the beam would be 4.6 dB *over a dipole* (which is sometimes written as 4.6 dBd).

20-19 SUMMARY

1. Standing waves on a transmission line can be measured by using voltage, current, power or impedances.
2. Antenna impedance can be measured by using a Wheatstone bridge or

an R-X noise bridge. Simple instruments, however, measure only the radiation resistance, and ignore the reactive component of impedance.

3. The *velocity factor* of transmission lines causes the physical length of a half-wave line to be shorter than the calculated length (e.g., 492/*F*).
4. Antennas can be compared with each other to find *relative* gain.

20-20 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

20-21 QUESTIONS

1. The impedance of an ideal antenna at resonance is equal to the____________ ____________.
2. List two important properties of a transmission line: ____________ ____________ and ____________.
3. The velocity of propagation in a transmission line is ____________ than the velocity in free space.
4. Describe in your own words the meaning of "standing waves" as it applies to radio transmission lines.
5. Standing waves create ____________ and ____________ nodes and antinodes at half-wavelength intervals along the line.
6. Name four parameters that could be used to measure SWR.
7. SWR is dependent upon transmission line length. True or false?
8. Draw a circuit for an R-X antenna impedance bridge.
9. Draw circuit for an R-X noise bridge.
10. List two techniques for trimming a transmission line to half wavelength.

20-22 PROBLEMS

1. An antenna shows a feed-point current of 1.6 amperes and a feed-point voltage of 26 volts. Find its impedance at resonance.
2. Calculate the physical length (in feet) of a half-wavelength radiator in free space if the frequency is 18 mHz.
3. Calculate a practical length in feet of an 18-mHz half-wave dipole near the earth's surface.

4. Calculate the length (in feet) of a quarter-wavelength 12-mHz vertical radiator (at the earth's surface).
5. Calculate the physical length in feet required to make a half-wavelength stub at 22 mHz out of foam-filled coaxial cable.
6. The voltage along a transmission line is measured, and it is found to peak at 58 volts, and then drops to a minimum of 31 volts at a point 9 feet away. Assume a perfect free-space transmission line to calculate (a) the SWR and (b) operating frequency of the transmitter used to excite the transmission line.
7. It is found that the forward power at a point in a transmission line is 100 watts, and the reflected power is 7 watts. Find the VSWR.
8. The VSWR is measured as 3:1, and the forward power is 480 watts. Calculate the reflected power.

chapter 21

RADIO RECEIVER MEASUREMENT AND ALIGNMENT

21-1 OBJECTIVES

1. To learn the fundamentals of radio receivers.
2. To learn the principal parameters of radio receivers.
3. To learn how to make the measurements of radio receiver performance.
4. To learn how to align a radio receiver.

21-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. Draw the block diagram for a simple superheterodyne receiver.
2. What is meant by receiver *sensitivity?*
3. Define receiver *selectivity* in your own words.
4. FM receivers use a ____________ sensitive demodulator.

21-3 RECEIVER BASICS

A radio or television receiver must intercept the signal transmitted in the air, and then *demodulate* it to recover the *information* impressed on the signal at the transmitter. AM and FM receivers are functionally the same, except at the demodulator circuit, so we will begin this discussion by considering an AM receiver.

The most elementary receiver is the crystal set, which consists of nothing more than a signal diode shunted across a pair of earphones, in the simplest case. But for at least the past sixty years more complex varieties have been used.

An early attempt at receiver design preceded the diode used to demodulate AM signals, with a cascade chain of tuned RF amplifiers. This design, called the TRF receiver, faded because of technical difficulties in keeping several RF amplifiers tuned to the same frequency over a wide tuning range.

The superheterodyne design of Fig. 21-1 has been in standard use for over fifty years, and it proves to be the design even today. In this type of receiver, the RF signal captured by the antenna is heterodyned to another (usually lower) frequency called the *intermediate frequency* (IF). Most of the receiver's total gain is in the IF amplifier.

Figure 21-1 shows the usual stages found in a superheterodyne receiver. The RF amplifier receives the weak signal from the antenna, and amplifies it before applying the signal to the mixer stage. The RF amplifier is a tuned stage, so it serves two functions other than amplification: It improves the receiver's ability to reject unwanted signals, and it prevents radiation of the local oscillator through the antenna.

The mixer stage and local oscillator (LO) are used to convert RF frequency signals to the IF frequency. The LO signal is heterodyned against the RF to produce the IF frequency. The mixer is the nonlinear element required for heterodyne action. The four frequencies found in the output of the mixer are the RF, LO, (RF + LO), and (RF − LO). Most receivers use the difference frequency between the LO and RF as the IF, so a tank circuit tuned to (RF − LO) will select only the mixer product designated as the IF and reject the others. In some receivers a combination stage called a *converter* replaces the mixer/LO. Such a stage merely uses a single active element (i.e., transistor) for both LO and mixer functions.

The IF amplifier is also tuned, but the tuning is not variable as it is in the RF and LO circuits; the IF remains tuned to one frequency (RF − LO). Since an amplifier that is tuned to but one frequency is easily neutralized, the IF amplifier

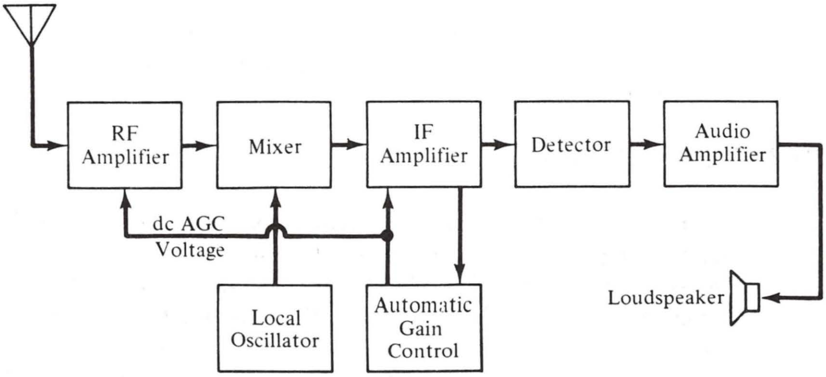


Figure 21-1. The superheterodyne receiver.

can provide most of the gain in the receiver without oscillating. Low-cost AM band broadcast receivers may have only one IF amplifier stage, while certain communications receivers and FM broadcast receivers have four or five stages in cascade.

The detector or demodulator stage recovers the information modulated onto the RF signal at the transmitter. In AM receivers, the detector is nothing more than a diode rectifier and filter circuit. Since this detector is sensitive to amplitude variations, it is called an *envelope detector*. The nonlinearity of the diode causes heterodyning between the AM signal's carrier and its sidebands, the difference frequency being the audio signal used to modulate the transmitter.

Single sideband (SSB) receivers must contend with the fact that the carrier was removed at the transmitter (Chapter 22), so is not available for heterodyning against a sideband. The SSB receiver, then, must use a *product detector* that consists of a mixer and fixed-frequency oscillator, followed by an audio frequency low-pass filter.

Frequency and phase modulation (FM and PM, respectively) must use a *phase-sensitive detector* to demodulate the carrier.

21-4 RECEIVER PARAMETERS

There are certain specifications that are common to all radio receivers, regardless of the frequency band that they cover. Some of these are *sensitivity, selectivity, image rejection,* and *noise figure*.

Sensitivity is a measure of the receiver's ability to pick up weak signals. Various definitions of sensitivity exist, leading to some confusion. One definition gives sensitivity as the level required across the antenna terminals to produce a certain amount of audio power at the output (usually 500 mW or 1 watt).

The other type of definition holds that sensitivity is the signal level at the input required to produce a 10- or 20-dB reduction in output level (the noise level compared with the noise level at zero signal level).

Selectivity is the ability to reject unwanted signals. A receiver does not look at the entire radio spectrum at one time, but only at a narrow window centered around the frequency indicated on the dial. The selectivity is defined in terms of bandwidth, so ideally the receiver selectivity matches exactly the bandwidth of the transmitted signal.

An *image* is a spurious response, and is due to the heterodyning process. The IF amplifier does not care whether the RF frequency is above or below the LO, because it accepts the difference. If the LO is above the RF, then the IF is (LO − RF). But a strong signal located at (LO + IF) also produces a difference frequency that is equal to the IF. The spurious signal is located at a frequency of twice the IF frequency from the RF, on the other side of the LO frequency, and the receiver must be able to reject that frequency.

Noise in a receiver (e.g., background hiss) is due largely to thermal agitation within the components. The noise figure in dB is the measure of how much noise is produced.

21-5 MEASURING SENSITIVITY

Recall that there are two definitions of sensitivity. In this section we will offer both methods, describing circumstances under which each is appropriate.

Figure 21-2 shows the equipment required for both types of sensitivity measurement. A calibrated-output signal generator is connected to the antenna terminals, and a voltmeter is connected across the output load. A load resistor *R* is used in place of the speaker; its value should be equal to the nominal loudspeaker impedance.

A dummy antenna, in receiver alignment or testing, is a passive network that simulates a transmission line of infinite length.

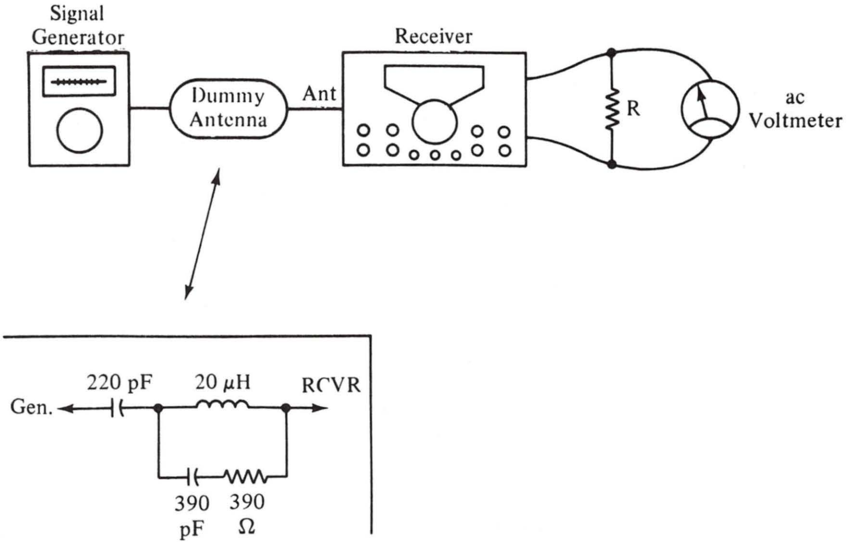


Figure 21-2. Measuring receiver sensitivity.

21-6 AUDIO POWER METHOD

The audio power method for measuring sensitivity is used on AM receivers. Most manufacturers recommending this technique specify an AM signal with a certain percentage of modulation between 25 and 100 percent (usually 30 percent at 400 or 1000 Hz).

Audio power is measured by noting the output voltage across the speaker load, and applying the relationship E^2/R. A typical manufacturer's specification will require 500 mW across an 8-ohm load, at 30 percent modulation of the carrier by a 400-Hz (or 1000-Hz) sine wave. This requirement means that we

1. Use a 400-Hz tone to modulate the generator 30 percent (Chapters 9 and 22).
2. Find the RF input signal in *microvolts* (μV) that will cause the output load to dissipate 500 mW, as indicated by an rms output voltage of

$$E = (PR)^{1/2}$$
$$E = [(0.5)(8)]^{1/2}$$
$$E = (4)^{1/2} = \mathbf{2\ volts}$$

3. We initially set the output control on the generator to zero, and then slowly advance the level higher until the output voltmeter reads 2 volts rms.

Some receivers have an *automatic gain control* (AGC) circuit that senses the signal level, and then creates a dc voltage proportional to the signal level. This voltage is fed back to the RF and IF stages to reduce their gain as signal strength increases. The AGC circuit must be disabled before any sensitivity measurement is possible. In some receivers, an AGC-ON-OFF switch is provided. This switch should be set to OFF. If there is no switch, then refer to the manufacturer's service manual for instructions on disabling the AGC. Some will have you disconnect it from the AGC rectifier, and clamp it to a given voltage (usually several volts negative).

21-7 THE QUIETING METHOD

The *quieting method* also uses the same equipment shown in Fig. 21-2. But the signal generator must be *unmodulated.*

1. Set the signal generator output control to zero, and turn up the receiver volume and gain controls to maximum.
2. Measure the rms voltage across the load produced by the noise.
3. Increase the input signal level until the output voltage drops 10 dB (some use 20 dB) from its zero signal level. If a properly calibrated dB voltmeter is not available, then rearrange the standard formula for voltage dB to solve for V_{final}.

$$\text{dB} = 20 \log_{10} \left[\frac{V_{\text{initial}}}{V_{\text{final}}}\right] \tag{21-1}$$

4. Read the sensitivity in microvolts from the output control of the signal generator.

Example 21-1

We want to measure the sensitivity of a receiver using the quieting method. It is required that we find out the signal level required to produce 10 dB of quieting. Assume that the zero signal output noise voltage is 5.6 volts rms, and calculate the voltage that should exist when the receiver is quieted −10 dB.

SOLUTION

$$\begin{aligned} 10 &= 20 \log_{10} (V_i/V_f) \qquad (21\text{-}1)\\ 10/20 &= \log_{10} (V_i/V_f)\\ 10^{(10/20)} &= V_i/V_f\\ 3.16 &= V_i/V_f\\ V_f &= V_i/3.16\\ V_f &= (5.6)/(3.16) = \mathbf{1.8\ volts\ rms} \end{aligned}$$

An alternate quieting method for AM receivers is to measure the signal level required to produce an output signal 10 dB above the noise level from a 30 percent modulated AM signal.

1. Connect the apparatus as in Fig. 21-2, set the signal generator output level to zero, and then adjust the receiver volume control to produce a readable output on the meter. If a calibrated audio VU or dB meter is available, then set the output level to 0 dB or 0 VU.
2. Set the signal generator modulation to 30 percent, and adjust the output level until the receiver output level increase 10 dB. [If a dB meter is available, then use Eq. (21-1).]
3. Read the sensitivity from the signal generator output level control, or meter. (Again, it is necessary to disable the AGC of the receiver.)

21-8 SELECTIVITY MEASUREMENT

Selectivity is the bandwidth of the receiver, and is the measure of its ability to reject unwanted signals. Most of the receiver's selectivity is due to the IF amplifier bandwidth; the RF and mixer bandwidth being somewhat broader. In most cases, therefore, selectivity is measured by injecting a signal at the IF frequency into the system at the output of the mixer (or alternatively, some manufacturers use the mixer RF input port to avoid detuning the first IF tank circuit). It is best to disable the LO regardless of which point is used for signal injection. Also, as before, disable the AGC circuit.

Figure 21-3 shows the equipment connection for selectivity measurement. The output indicator will be either a dc voltmeter across the AGC line, or an RF voltmeter (i.e., a dc voltmeter with an RF detector probe) immediately preceding the receiver detector. If the receiver is an AM model, then an ac voltmeter across the speaker load may also be used. In the latter instance a modulated signal generator is needed.

1. Adjust the signal generator to a convenient output level, and then tune it to the IF frequency, as indicated by a maximum output on the indicator.
2. Note the output voltage and the generator frequency. (*Note:* The output voltage obtained in Step 2 is used as the 0-dB reference level.)
3. Measure the output voltage at equally spaced frequencies above and below the IF, and record the voltages on a graph.
4. Plot the data *dB down-vs-ΔF* in the form shown in Fig. 21-4(b). The selectivity is usually *defined* as the bandwidth in hertz or kilohertz between points where the response is −6 dB from the response at the IF frequency.

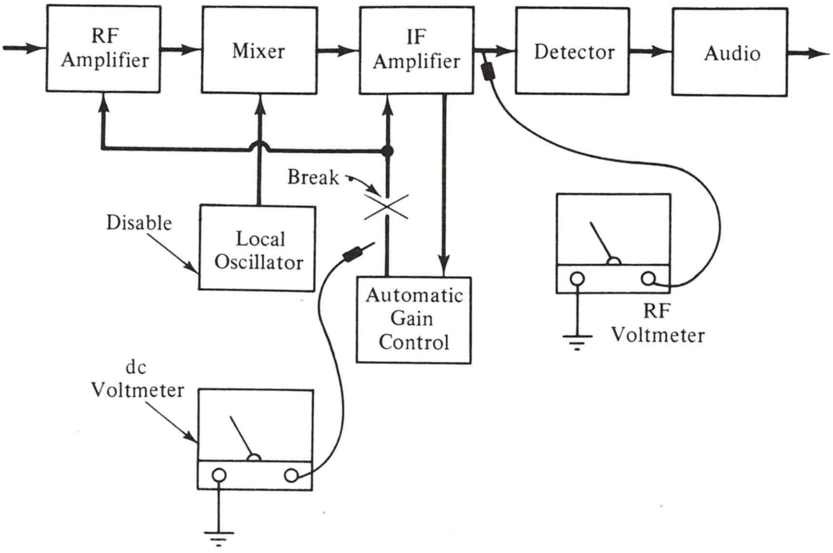


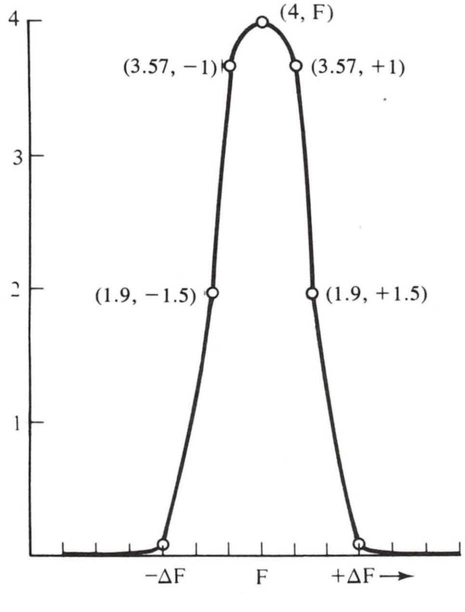
Figure 21-3. Measuring selectivity.

The selectivity of receivers may be different in different reception modes, or there may be a means for varying the selectivity as needed. In any event, the technique above is terribly time-consuming. A superior technique is the sweep method of the next section.

$-\Delta F$	E	dB	$+\Delta F$	E	dB
0	4	0*	+1	3.57	−1
−1	3.57	−1	+1.5	1.9	−6.5
−1.5	1.9	−6.5	+2	0.25	−24
−2	0.25	−24	+3	0.016	−48
−3	0.016	−48	+4	0.010	−52
−4	0.010	−52	+5	0.007	−55
−5	0.007	−55			

*By definition

(a)



(b)

Figure 21-4 (a) Trial data. (b) Data plotted.

21-9 SWEEP METHOD

In the sweep method, a sweep or FM signal generator is used to sweep back and forth across the IF frequency. Figure 21-5(a) shows an appropriate equipment connection scheme. In some cases, incidentally, the sweep generator signal is at the RF frequency and is applied to the antenna terminals of the receiver.

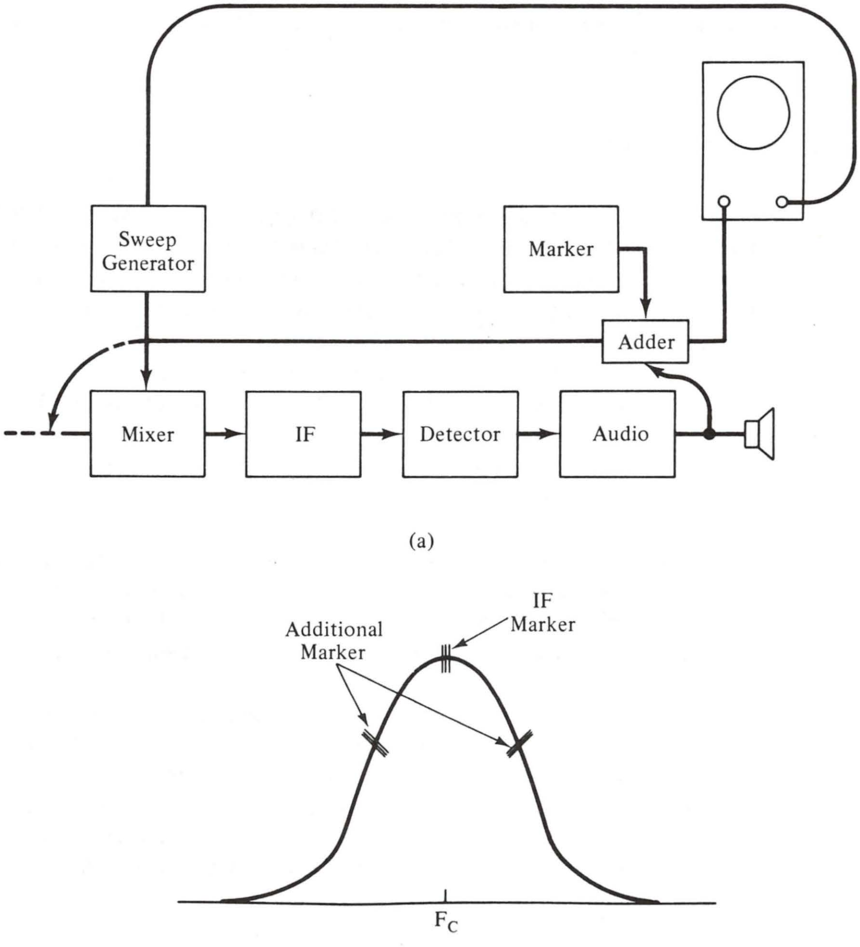


Figure 21-5. (a) Sweep alignment equipment setup. (b) Sweep alignment IF response curve.

The audio output of the receiver, the marker, and sweep signals are all applied to an *adder* before going to the vertical input of the oscilloscope. The marker oscillator(s) superimposes "pips" on the waveform [Fig. 21-5(b)] indicating precise frequency points.

In the sweep method, we generally use the markers and the horizontal gain control of the oscilloscope to calibrate the oscilloscope screen in kHz/div. This calibration allows us to read bandwidth directly from the horizontal scale. The −6-dB points, incidentally, are the points where the *voltage* response drops to one-half of the output level at the IF frequency. These points are also known as the *half-power points*.

21-10 IMAGE RESPONSE

In Sec. 21-4 we defined the problem of image response in superheterodyne receivers. One of the most notable differences between poor and good radio receivers is this parameter. In fact, low-cost shortwave receivers are so poor in this respect that it is often difficult to discern which is the signal and which response is the image, especially in the over-15-mHz band.

Image response is reduced by using a sharply tuned, high-Q, RF amplifier "front end," and a high IF frequency. A 9-mHz IF, for example, produces images 2×9 mHz = 18 mHz away from the RF, but a 455-kHz IF produces images only 0.91 mHz away from the RF signal. The measurement of the image response is as follows:

1. Tune the receiver and the signal generator to the same frequency. Adjust the generator output control for 50 μV, or some setting that gives a readable output voltage across the load. Call this voltage $V1$.
2. Record the output voltage and signal generator attenuator setting (i.e., $V1$).
3. Tune the signal generator to a frequency equal to 2 × IF above or below the RF frequency, until the signal is heard in the output. Fine-tune the signal generator (*not* the receiver) for maximum output.
4. Adjust the generator output attenuator for the same receiver output as was obtained in Step 1. The generator output that creates this level, in microvolts, is $V2$. Apply the standard voltage decibel equation to find the image rejection in dB:

$$\text{dB} = 20 \log_{10} (V2/V1)$$

The result obtained is now the image response of the receiver in dB. Note that the image may occur above or below the initial RF signal, depending upon whether the LO is above or below the RF frequency.

21-11 AM ALIGNMENT

The alignment of simple AM receivers uses an equipment setup such as Fig. 21-6. Adjust the signal generator to produce a 30 percent modulation at a signal level that gives a readable output. Set the signal generator dial to a precisely known frequency within the receiver range. Set the receiver dial to the same point.

1. Adjust the local oscillator trimmer controls for maximum output signal.
2. Adjust the RF amplifier trimmer controls for maximum output signal.
3. Adjust the IF transformer tuning for maximum output signal.
4. Reduce the signal generator output level until noise is heard in the receiver output.
5. Repeat Steps 1 through 4 until no further improvement is possible.

The procedure given above is a generalization that will usually work on most receivers. Some manufacturers, however, specify their own procedure; especially on high-cost shortwave receivers.

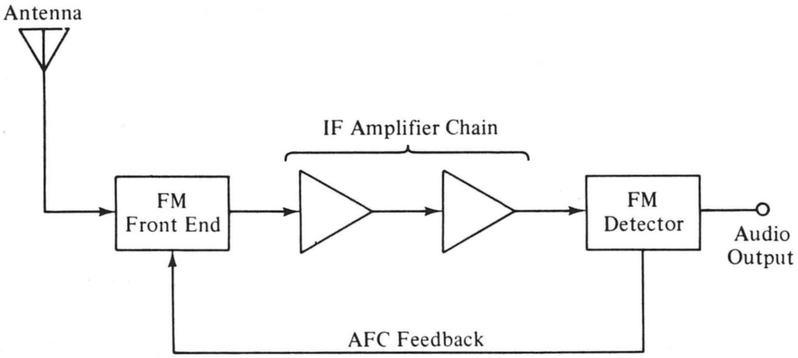


Figure 21-6. Block diagram of an FM receiver.

21-12 FM RECEIVER ALIGNMENT

The FM receiver is similar to the AM receiver, except in the demodulator. An FM receiver requires a phase-sensitive detector, whereas an AM receiver uses an envelope detector. Also, most broadcast and two-way communications FM receivers use three to five stages of IF amplification. Some models of more recent design use one or two integrated circuit gain blocks.

Many communications FM or PM receivers are dual conversion models such as that shown in Fig. 21-7. In a dual conversion receiver, the front end converts the RF signal to some high IF frequency, usually 10.7 mHz. This high IF signal is then heterodyned against the output of a crystal oscillator to produce a low IF frequency, usually 455 kHz.

One problem in alignment of FM receivers is to determine which of several popular FM detector circuits are used in the demodulator. Several types are in general use: *discriminator, ratio detector,* and *quadrature detector*.

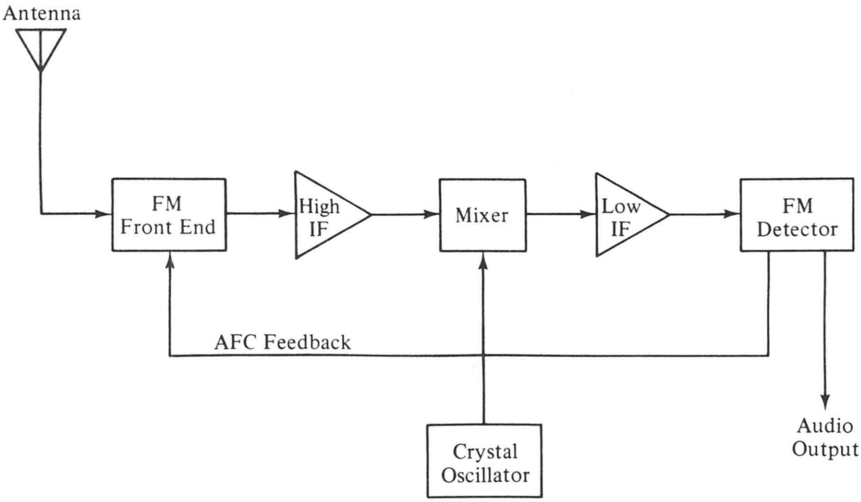


Figure 21-7. Block diagram of a dual conversion FM receiver.

Figure 21-8 shows the Foster-Seeley FM discriminator. In this circuit, signal voltages from $T1$ are added to signal voltages in the secondary. When the signal frequency is exactly equal to the frequency to which the secondary is tuned, the output voltage will be zero. FM deviation, whether caused by a shift in carrier frequency or by the process of frequency modulation, will produce a positive or negative output voltage, depending upon the direction of frequency shift.

The ratio detector is shown in Fig. 21-9. Note that there are two aspects of this circuit that are different from the discriminator: The diodes point in opposite directions, and the circuit includes an electrolytic capacitor ($C3$). This capacitor has, as one of its functions, the suppression of amplitude modulation or noise signals. Because of $C3$, receivers using a ratio detector often do not have limiter amplifiers.

In a ratio detector the relative charges on capacitor $C1$ and $C2$ will have a ratio of 2:1 when the IF frequency is precisely equal to the resonant frequency of the secondary of the transformer. Frequency deviation causes the ratio of the

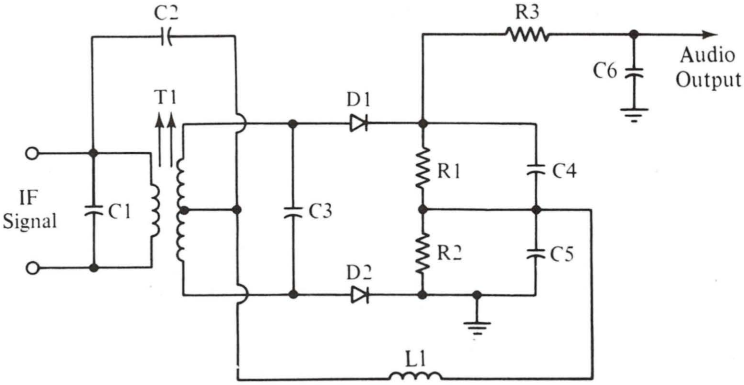


Figure 21-8. FM discriminator.

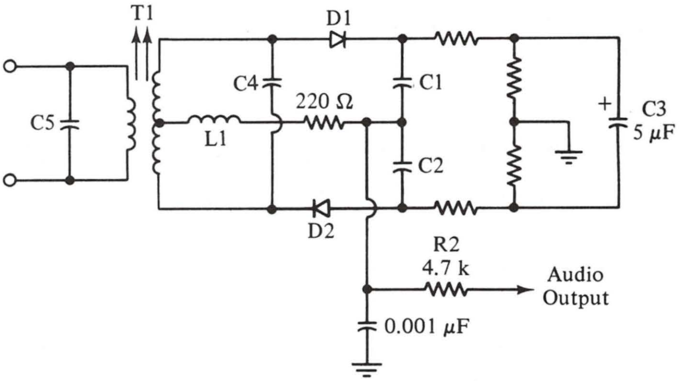


Figure 21-9. Ratio detector.

voltages across these capacitors to change, and this results in the audio output signal.

The last type of detector to be considered here is the quadrature detector. Although this type of detector was once popular in vacuum tube form, there are several IC models available. The name of this detector is derived from the fact that two components of the IF signal, phased 90 deg apart, are combined in a synchronous detector circuit. A block diagram for a quadrature detector is shown in Fig. 21-10. The input stages are a wideband (i.e., untuned) high-gain limiter amplifier that is used to eliminate any AM or noise components on the IF signal. The output of this limiter consists of square waves, which are fed to a signal splitter. One output of the splitter is applied directly to one input of the synchron-

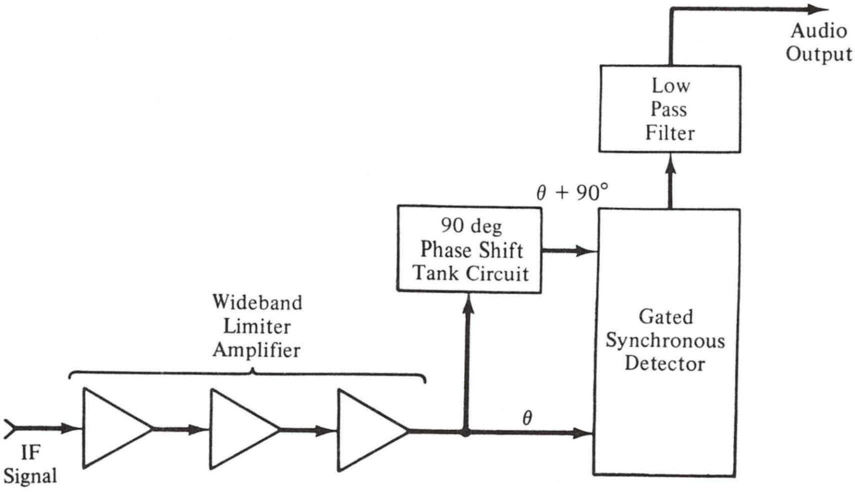


Figure 21-10. Block diagram to an IC quadrature detector.

ous detector, while the other is passed through a tank circuit that shifts its phase 90 deg. The output of the detector is integrated to recover the modulation audio.

21-13 SWEEP ALIGNMENT

The test equipment configuration for sweep alignment is shown in Fig. 21-11. The FM signal generator signal, a precise crystal marker generator, and the receiver audio output are summed in an adder circuit before being applied to the vertical input of an oscilloscope. The RF signal is also applied to the antenna terminals of the receiver.

Figure 21-12 shows the curves that can be expected on the screen of the oscilloscope. The curve in Fig. 21-12(a) shows the IF frequency response curve. This curve would be found by applying the IF output signal to the oscilloscope through an RF voltmeter probe. The S-curve of Fig. 21-12(b) is the curve of the discriminator. The discriminator curve will pass through zero exactly at the IF center frequency when the detector transformer secondary is properly adjusted.

In some alignment procedures, a zero-center dc voltmeter is connected to the audio output line from the detector. The voltmeter will indicate zero output when the transformer is adjusted properly.

The quadrature detector is adjusted by using an ac voltmeter across the output of the receiver. The 90-deg phase shift network is then adjusted for maximum audio output.

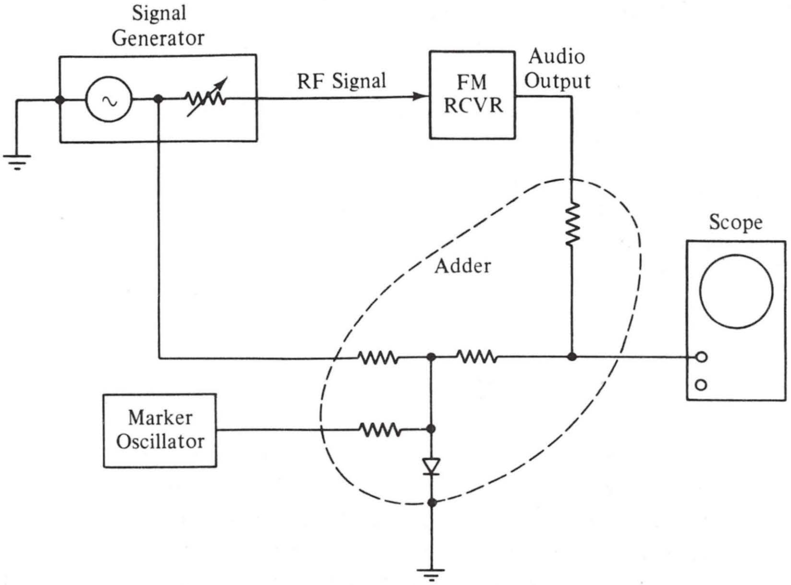


Figure 21-11. FM sweep alignment test equipment setup.

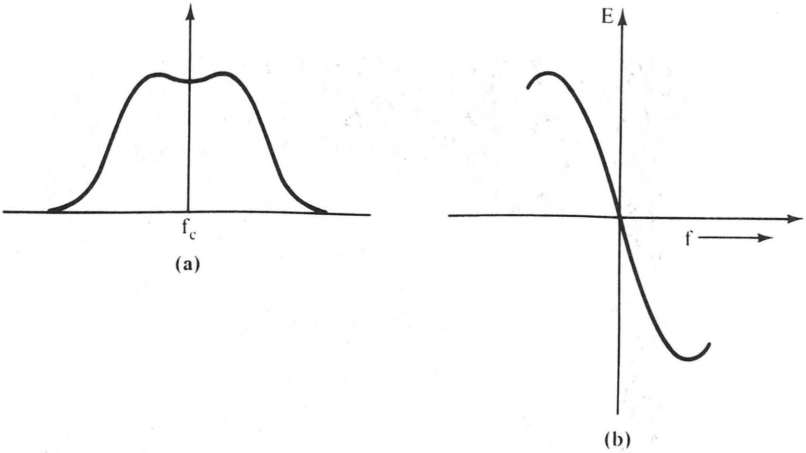


Figure 21-12. (a) FM IF response curve (dip should be no more than 10 percent of trace amplitude). (b) Discriminator "S-curve" response.

Any type of FM receiver can be adjusted properly by using a total harmonic distortion analyzer (THD) at the receiver output. Adjust the FM detector transformer for minimum THD reading.

21-14 NONSWEPT ALIGNMENT

If only nonswept signal generators are available, it is still possible to perform the alignment. Connect the RF output of the signal generator to the antenna terminals of the receiver, and connect a zero-center dc voltmeter across the audio output of the detector, not the receiver. The proper connection point is prior to any ac coupling that is used. Set the signal generator to the precise center IF frequency, and then adjust the detector transformer for a zero volts indication. This technique is used primarily on discriminator detectors.

21-15 DUAL-SWEEP ALIGNMENT

The Sound Technology Model 1000A signal generator (Chapter 9) uses the dual-sweep technique for FM alignment. In this method, a low-frequency (i.e., 60-Hz) signal and a 10-kHz signal are used to modulate the FM oscillator. With proper oscilloscope connections, the output trace will appear as in Fig. 21-13. This trace contains points that are proportional to the intermodulation distortion of the output signal, and this is directly related to the state of the IF alignment. The detector is adjusted for maximum flatness of the curve.

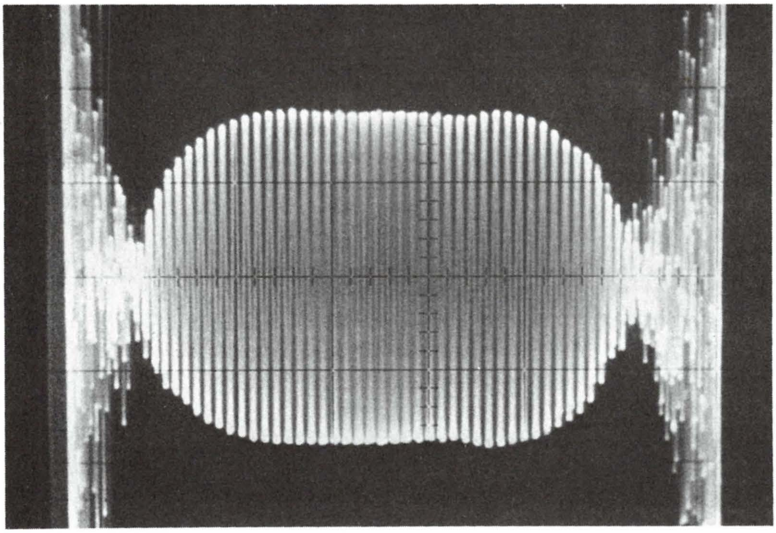


Figure 21-13. Dual-sweep trace. (Courtesy of Sound Technology, Inc.)

21-16 SUMMARY

1. Four important receiver parameters are sensitivity, selectivity, image rejection ratio, and noise figure.
2. AM receivers use an envelope detector; SSB receivers use a product detector; and FM/PM receivers use a phase-sensitive detector.
3. Receiver sensitivity can be defined in terms of signal required to produce a given audio output, or level required to produce a 10- or 20-dB quieting level.
4. Selectivity is usually measured as the bandwidth between −6 dB-points on the IF response curve.

21-17 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a check mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

21-18 QUESTIONS

1. Define *selectivity* in your own words.
2. Define *sensitivity* in your own words.
3. Describe two methods for measuring sensitivity.
4. Describe a method for measuring selectivity.
5. There is more than one method for specifying and measuring sensitivity. Do these two methods always yield the same numbers?
6. List three types of FM detector circuit.
7. In an FM detector, the output voltage will be ____________ when the input signal is unmodulated and precisely tuned to the IF center frequency.
8. In most receiver measurements or alignment procedures the ____________ should be disabled.
9. The ____________ stages of a receiver provide most of the gain and selectivity.
10. Spurious response to a signal removed from the RF signal by a factor of 2 × IF results from poor ____________ rejection.

21-19 PROBLEMS

1. A receiver being tested for sensitivity by the audio power method uses a 10-ohm speaker. Calculate the output voltage that indicates a 500-mW output power level.
2. We are testing receiver sensitivity by the quieting method using an unmodulated signal generator. If the zero signal noise level is 6.9 volts rms, what level represents 10 dB of quieting?

chapter 22

RADIO TRANSMITTER MEASUREMENTS

22-1 OBJECTIVES

1. To learn how RF power is measured.
2. To learn the different types of transmitter monitor instrument available.
3. To learn how AM and FM modulation is measured.
4. To review frequency measurement techniques.

22-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer any particular question, place a check mark beside it, and then look for the answer as you read the text.

1. List two different types of RF wattmeter.
2. A ____________ ____________ meter consists of a discriminator and an output meter that is calibrated in units of frequency.
3. SSB transmitters are usually tested with a ____________ ____________ audio signal.
4. Thermocouple RF ammeters can be used to measure RF power up to approximately ____________ mHz.
5. An ____________ oscilloscope can be used as a linearity indicator for RF linear amplifiers.

22-3 TRANSMITTER MEASUREMENTS

It is necessary to measure certain parameters on radio transmitters. Some measurements are required to assure that the transmitter meets FCC specifications and regulations. In other cases, measurements are made to repair the equipment, or optimize the performance of newly installed transmitters. In still other cases manufacturers must make certain types of measurements to confirm engineering designs, or to correct defects discovered in the testing.

In this chapter, we will consider certain measurements that have been considered in other chapters. You are advised, however, that it might be prudent to review Chapters 17, 19, and 21, and those sections of Chapter 9 that deal with frequency synthesizers.

22-4 RF POWER MEASUREMENT

Power measurements at RF frequencies are sometimes considered difficult to achieve with any accuracy. But even crude, home-built instruments will yield ±25 percent accuracy, while certain low-cost professional equipment will yield better than ±5 percent accuracy.

Figure 22-1 shows a method often used in AM broadcasting stations, in which a thermocouple RF ammeter is connected in series with the load (i.e., antenna or a dummy load). Such meters are inherently rms-reading devices, so will measure the RF power from a wide variety of RF waveforms. RF power can be calculated from the conventional formula

$$P = I^2 R_L \tag{22-1}$$

where P = the powered delivered to the load
I = the line current to the load
R_L = the resistance of the load

Example 22-1

Calculate the power delivered to a 75-ohm load, if the line current is measured as 1.8 amperes rms.

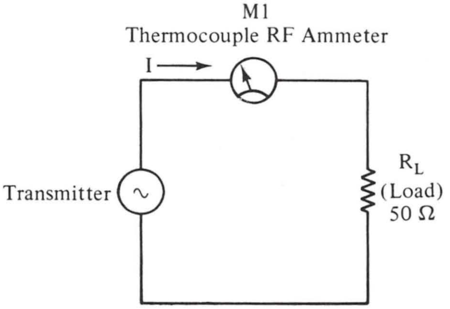


Figure 22-1. Measuring RF power with a thermocouple RF ammeter.

SOLUTION

$$P = I^2 R_L \tag{22-1}$$
$$P = (1.8\ \text{A})^2(75\ \text{ohms})$$
$$P = (3.24\ A^2)(75\ \text{ohms}) = \mathbf{243\ watts}$$

(Verification that the units $A^2\ \Omega$ are equivalent to watts is left as an exercise for the reader.)

Radio frequency ammeters produce usable results at frequencies up to approximately 50 mHz, or so. Most of these instruments are labeled "frequency compensated to 50 mHz," or some frequency in that region.

RF voltmeters are also used to make power measurements, although few of these instruments are actually rms-reading devices. Figure 22-2 shows an RF voltmeter that could be used to make RF power measurements. Resistor $R1$ is the load, while resistors $R2$ and $R3$ form a *voltage divider*. The voltage divider output is the voltage drop across $R3$, which is rectified by $D1$, and is used to charge $C2$ to the waveform's peak voltage. The rms voltage is equal to 0.707 times the peak voltage. The voltmeter, which is actually calibrated in watts instead of volts, measures a potential of

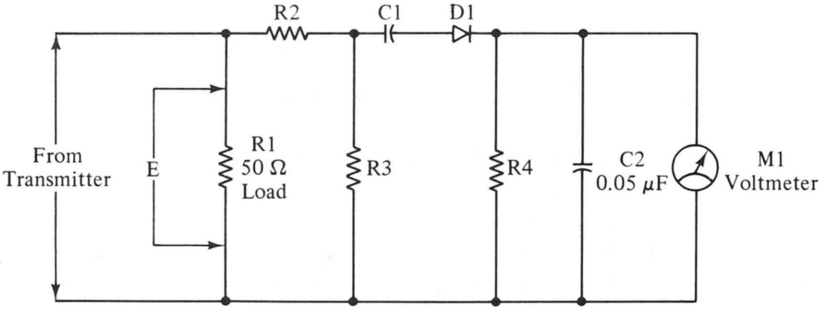


Figure 22-2. Measuring RF power by the voltage across the load.

$$E_m = \frac{ER3}{(R2 + R3)} \tag{22-2}$$

where E_m = the voltage across the meter
E = the peak voltage across the load
$R2$ = the value of R2
$R3$ = the value of R3

To find the RF power, we must first rewrite Eq. (22-2) to solve for *E:*

$$E = \frac{E_m(R2 + R3)}{R3} \tag{22-3}$$

Once *E* is determined, we may use the expression

$$P_p = \frac{E^2}{R1} \tag{22-4}$$

to calculate the RF power dissipated in load resistor *R*1. But care must be used when measuring power with this type of circuit, because the RF voltmeter is a peak-reading device. To find rms power, which is the important measurement, we use the expression

$$P = \frac{(0.707\,E)^2}{R1} \tag{22-5}$$

But this expression holds true only if the RF power waveform is sinusoidal. If any other waveform is used, then a correction factor other than 0.707 must be used, or the error must be tolerated.

An RF wattmeter using a toroidal core current transformer is shown in Fig. 22-3. This same type of instrument is also used in SWR measurements (Chapter 20). This type of meter also suffers from the necessity of ensuring a sinusoidal RF waveform.

Capacitors *C*1 and *C*2 are used to frequency-compensate the instrument so that it is useful over a wide frequency range, i.e., 2 to 30 mHz.

There is no simple relationship between the RF power and the indication given by meter *M*1. Calibration, therefore, is best done by using an RF ammeter to correlate power and *M*1 readings.

The example shown in Fig. 22-2 showed a resistor voltage divider, but capacitor, inductor, and RC/RL dividers are also seen occasionally.

22-5 THERMAL METHOD

When any electrical energy is dissipated in a resistance, heat is created. In fact, some textbooks define rms electrical values in terms of the amount of dc power required to generate the same amount of heat in a standard resistance. This same phenomenon gives us a technique for thermal RF power measurements.

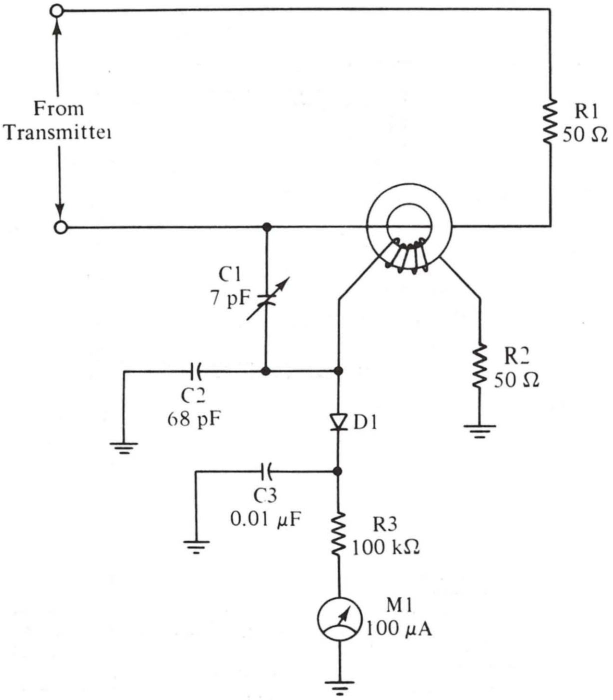


Figure 22-3. Toroid current transformer-type RF wattmeter.

Figure 22-4(a) shows the basic requirements for such an instrument. The load is not the antenna, but a *dummy load,* i.e., a standard-value, noninductive resistor. A thermal sensor is closely coupled to the load, and both are inside a thermal chamber that isolates them from outside heat sources.

An early version of this system placed the load inside an oil-filled, thermally insulated chamber, and a mercury thermometer measuring the oil temperature was visible from the outside of the chamber. The operator noted the temperature before and after the power was applied, and then used a nomograph to calculate the power dissipated from the temperature data.

But this method is too cumbersome, because it is necessary to allow the oil temperature to come to thermal equilibrium after RF power is applied. This procedure might take several minutes. Additionally, there is no direct power indication, so the user has to use a graph or equation to interpret the results.

Modern thermal RF power meters use faster thermometry elements to make the measurement. All of the electronic thermal transducers of Chapter 13 have been used in this type of application. Being electronic, these thermometers can be calibrated directly in power units, i.e. watts.

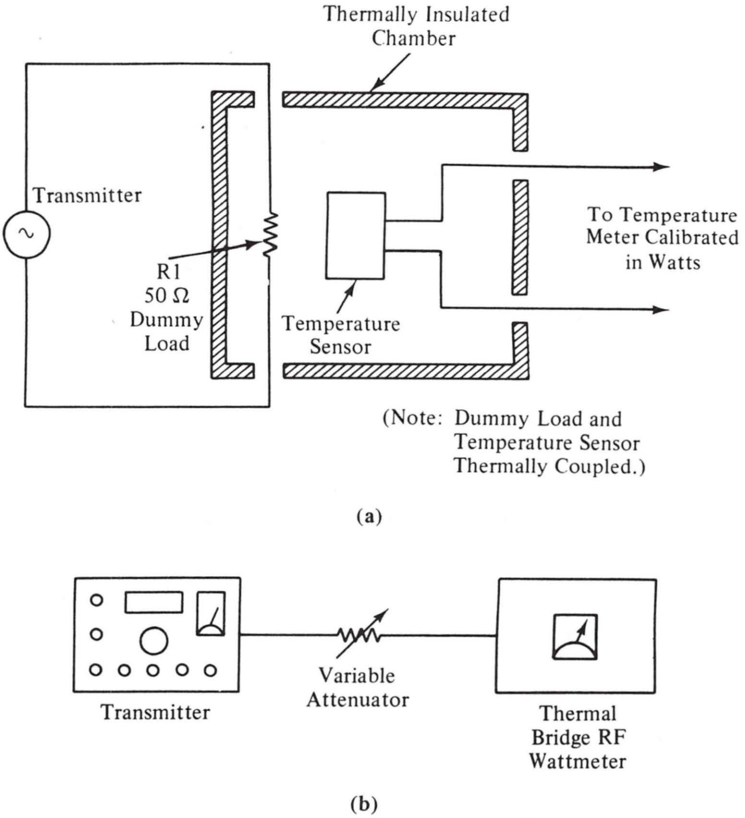

Thermally Insulated Chamber
Transmitter
R1
50 Ω
Dummy
Load
Temperature Sensor
To Temperature Meter Calibrated in Watts
(Note: Dummy Load and Temperature Sensor Thermally Coupled.)
(a)
Transmitter
Variable Attenuator
Thermal Bridge RF Wattmeter
(b)

Figure 22-4. (a) Thermal method for measuring RF power. (b) Test equipment setup for measuring RF power using the thermal method.

Thermal sensors have the advantage of being inherently rms indicating, and will operate over a wide frequency range. One instrument, for example, boasts a frequency response from dc to the microwave range. The limitation on frequency response is a function of the properties of the dummy load.

Some thermal RF power meters have only a low power measuring range. One, for example, will measure RF power levels up to 100 milliwatts. We can extend the power range by using a calibrated attenuator [Fig. 22-4(b)] between the transmitter and the thermal meter. The attenuator may be a precision step-attenuator that can be adjusted, or it may be a fixed "barrel" attenuator. The latter type of attenuator uses a fixed resistance network inside an extended RF connector. Attenuation factors are given in dB, which is expressed by

$$\text{dB} = 10 \log_{10} \frac{P1}{P2} \tag{22-6}$$

where dB is the notation for decibels
$\log_{10}$ denotes the base-10 logarithms
$P1$ is one power level, i.e., the input to the attenuator
$P2$ is the other power level, i.e., the attenuator output

We may measure higher full-scale power levels by setting $P2$ equal to the wattmeter reading, and then solving Eq. (22-6) for $P1$ by taking 10 to a power equal to both sides of the equation:

$$\text{dB} = 10 \log_{10} \frac{P1}{P2}$$

$$\frac{\text{dB}}{10} = \log_{10} \frac{P1}{P2}$$

$$10^{(\text{dB}/10)} = P1/P2$$

$$P1 = (P2)(10^{(\text{dB}/10)}) \tag{22-7}$$

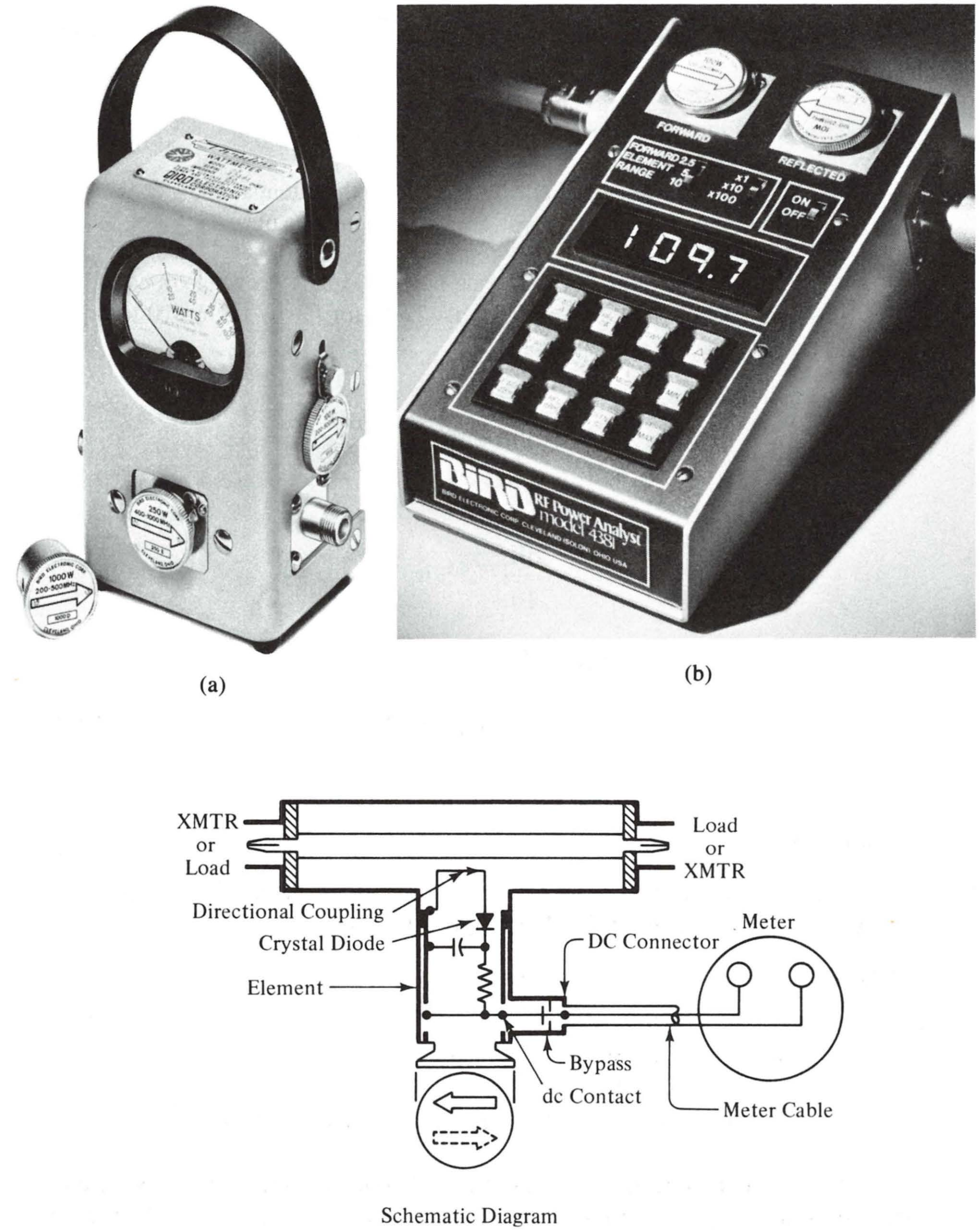
Example 22-2

A thermal RF wattmeter connected as in Fig. 22-4(b) reads 84 mW when 15 dB of attenuation is used. Find the applied power.

SOLUTION

$$P1 = (P2)(10^{(\text{dB}/10)}) \tag{22-7}$$
$$P1 = (84\ \text{mW})(10^{(15/10)})$$
$$P1 = (84\ \text{mW})(10^{1.5})$$
$$P1 = (84\ \text{mW})(32) = 2688\ \text{mW} = \mathbf{2.688\ watts}$$

The thermal RF wattmeter is used with a dummy load, so it will be able to measure transmitter output power. But to measure the RF power applied to an antenna system or other active load, we must use an RF wattmeter that may be inserted in the transmission line. Simple instruments based on circuits such as Figs. 22-2 and 22-3 will work only under the right conditions.



(a)

(b)

Schematic Diagram

(c)

Figure 22-5. (a)Bird Model 43 Thruline® RF Wattmeter. (Courtesy of Bird Electronics Corp.) (b) Digital readout Thruline® RF Wattmeter. (Courtesy of Bird Electronics Corp.) (c) Construction of the Thurline® RF Wattmeter. (Courtesy of Bird Electronics Corp.)

22-6 THRULINE®* WATTMETERS

The Thruline RF wattmeter can be inserted into the transmission line of an antenna system, and presents so little loss that it may be left inserted during normal transmitter operations. Figure 22-5(a) and 22-5(b) show examples of the Thruline RF wattmeters that are popular with two-way radio servicers. The Thruline RF wattmeters use a directional coupler assembly, shown in Fig. 22-5(c), connected in series with the transmission line. The plug-in directional element can be rotated 180 degrees to measure both forward and reflected power levels. A sampling loop and diode detector are contained within each plug-in element. The main RF barrel is actually a special coaxial line segment with a 50-ohm characteristic impedance. The Thruline® sensor works due to the mutual inductance *M* between the sample loop and the center conductor of the coaxial element. Figure 22-6 shows an equivalent circuit. The output voltage from the sampler *(e)* is the sum of two voltages, e_r and e_m.

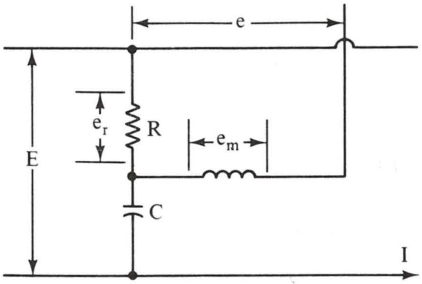


Figure 22-6. Equivalent circuit for Thruline® RF wattmeter.

Voltage e_r is created by the voltage divider action of *R* and *C* on transmission line voltage *E*. If *R* is much less than X_c, then we may write the expression for e_r below:

$$e_r = \frac{RE}{X_c} = RE(j\omega C) \tag{22-8}$$

Voltage e_m, on the other hand, is due to mutual induction, and is expressed by

$$e_m = I(j\omega) \pm M \tag{22-9}$$

We now have the expression for both factors contributing to output voltage *e*. We know that

---

*"Thruline®" is a registered trademark of Bird Electronics.

$$e = e_{\mathrm{r}} + e_{\mathrm{m}} \tag{22-10}$$

So, by substitution,

$$e = j\omega(CRE \pm M) \tag{22-11}$$

If the sample element components are selected so that

$$CR = M/Z_{\mathrm{o}} \tag{22-12}$$

we may then write Eq. (22-11) in the form

$$e = j\omega M[(E/Z_{\mathrm{o}}) \pm I]$$

by recognizing that, at any given point in a transmission line, E is the sum of the forward (E_{f}) and reflected (E_{r}) voltages, and that the line current is equal to

$$I = \frac{E_{\mathrm{f}}}{Z_{\mathrm{o}}} - \frac{E_{\mathrm{r}}}{Z_{\mathrm{o}}} \tag{22-14}$$

We may specify e in the forms

$$e = \frac{j\omega M(2E_{\mathrm{f}})}{Z_{\mathrm{o}}} \tag{22-15}$$

and

$$e = \frac{j\omega M(2E_{\mathrm{r}})}{Z_{0}} \tag{22-16}$$

The output voltage e, then, is proportional to the mutual inductance and frequency. But the manufacturer terminates e in a capacitive reactance, so the frequency dependence is lessened (see Fig. 22-7). Each element is custom compensated, therefore, for a specific frequency and power range. Beyond the specified range for any given element, accuracy is not guaranteed.

The basic Thruline RF wattmeter is valid only in CW and unmodulated AM/FM systems. In many systems, however, peak power readings may be required. Figure 22-8 shows the envelope waveforms for several different types of transmitter emission. For some of these waveforms, a peak-reading RF wattmeter is required. Bird Electronics makes the model 4311, based on their basic model 43 system, in which an IC amplifier-and-servo system finds the peak envelope voltage and uses it to derive the peak power value.

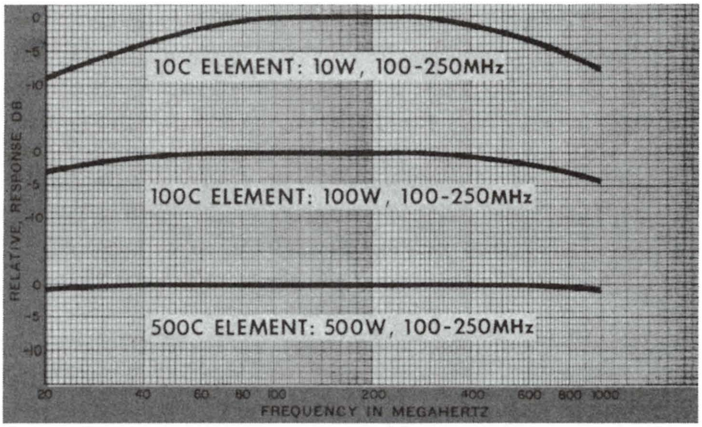


Figure 22-7. Frequency response of various Bird Model 43 elements. (Courtesy of Bird Electronics Corp.)

Consider an amplitude-modulated (AM) transmitter that is modulated 100 percent by a sinusoidal audio signal. If the transmitter is able to produce 100 watts of unmodulated CW power, it produces a peak power in the modulation envelope of 400 watts (i.e., $P_p = 4 \times P_a$).

22-7 DUMMY LOADS

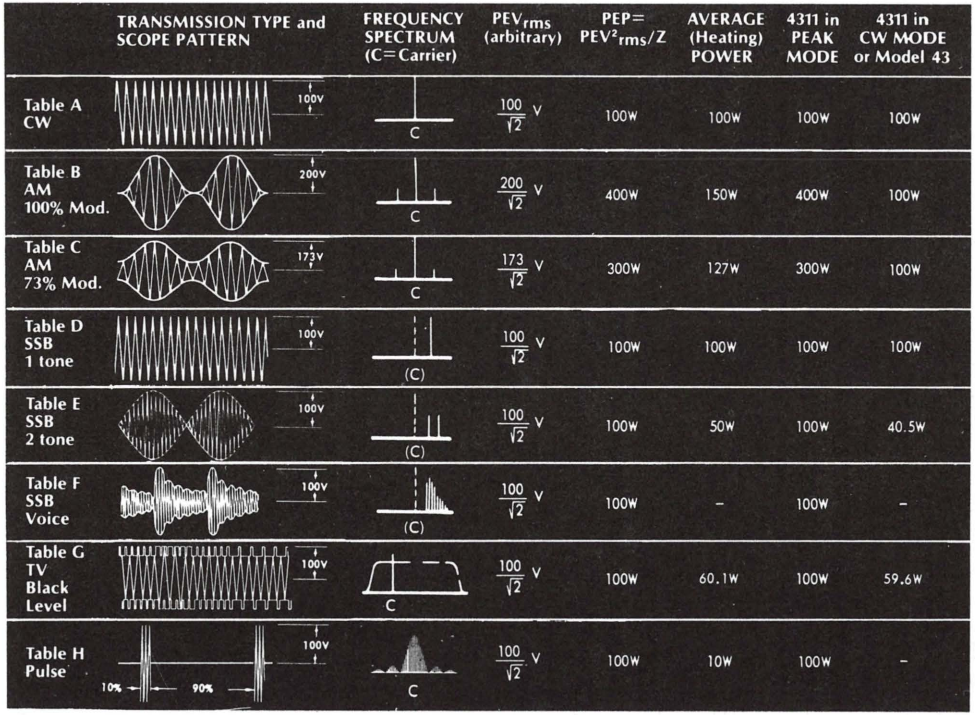
It is illegal to perform some types of transmitter measurement with a live (i.e., radiating) antenna, because of possible interference to other stations. In other cases, it is simply not advisable to use an antenna, so *dummy loads* are used to absorb transmitter power.

A dummy load is a noninductive, pure resistance inside a shielded metal container (see Fig. 22-9). The value of the resistor is selected to match the impedance of the transmitter. Since most commercial antenna systems are designed to match 50 ohms, the majority of the dummy loads on the market have a value of 50 ohms.

An instrument called an absorption wattmeter is made by pairing a dummy load and an RF wattmeter. These instruments are often used on the test bench in making transmitter measurements, or in certain applications in the broadcast industry.

22-8 MODULATION MEASUREMENTS

Modulation is an information-carrying signal that modifies a radio carrier signal. The modulation may consist of audio speech signals or tones that represent analog or digital data.



	TRANSMISSION TYPE and SCOPE PATTERN	FREQUENCY SPECTRUM (C=Carrier)	PEV_{rms} (arbitrary)	PEP= PEV^2_{rms}/Z	AVERAGE (Heating) POWER	4311 in PEAK MODE	4311 in CW MODE or Model 43
Table A CW	100V	C	$\frac{100}{\sqrt{2}}$ V	100W	100W	100W	100W
Table B AM 100% Mod.	200V	C	$\frac{200}{\sqrt{2}}$ V	400W	150W	400W	100W
Table C AM 73% Mod.	173V	C	$\frac{173}{\sqrt{2}}$ V	300W	127W	300W	100W
Table D SSB 1 tone	100V	(C)	$\frac{100}{\sqrt{2}}$ V	100W	100W	100W	100W
Table E SSB 2 tone	100V	(C)	$\frac{100}{\sqrt{2}}$ V	100W	50W	100W	40.5W
Table F SSB Voice	100V	(C)	$\frac{100}{\sqrt{2}}$ V	100W	-	100W	-
Table G TV Black Level	100V	C	$\frac{100}{\sqrt{2}}$ V	100W	60.1W	100W	59.6W
Table H Pulse	100V 10% 90%	C	$\frac{100}{\sqrt{2}}$ V	100W	10W	100W	-

In the table above, $Z_0=50$ ohms, PEP is Peak Envelope Power, and PEV is Peak Envelope Voltage. The PEV of the Carrier (or suppressed Carrier) C was arbitrarily chosen at 100 volts in all examples. $PEV_{rms}=PEV/\sqrt{2}$

Figure 22-8. (Courtesy of Bird Electronics Corp.)

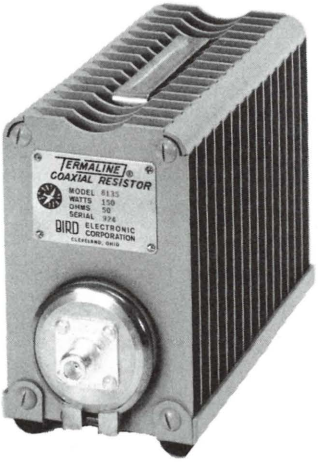


Figure 22-9. Dummy load. (Courtesy of Bird Electronics Corp.)

In most communications applications one or more of three basic properties of the carrier signal are varied by the information signal: *amplitude, frequency,* or *phase*. We will consider amplitude modulation (AM) in one class, and in a class together frequency (FM) and phase modulation (PM).

22-9 FM DEVIATION METERS

The relationship between the amplitude of the audio signal and the FM transmitter's output frequency is shown in Fig. 22-10(a). When the signal amplitude is zero, or when there is no signal present, the output frequency is F_0; this frequency is the assigned carrier frequency of the transmitter. When the audio signal amplitude begins increasing in the positive direction, the carrier frequency increases to F_h. Similarly, on negative excursions of the audio signal, the carrier frequency decreases to F_L. The measure of frequency modulation is *deviation,* which is defined as the frequency *change* (ΔF) from F_0 to either F_h or F_L. Positive deviation, then, is the quantity $(F_h - F_0)$, and negative deviation is $(F_0 - F_L)$.

Figure 22-10(b) shows a typical deviation meter in block diagram form. It is basically an FM superheterodyne receiver with a dc voltmeter across the output of the discriminator detector. The *voltage-vs-frequency* characteristic of a discriminator is the S-shaped curve of Fig. 22-10(c). The voltage on one side of the discriminator circuit will be positive an amount proportional to the positive deviation, while the voltage on the other half is negative an amount proportional to the negative deviation. A commercial example is shown in Fig. 22-10(d).

An FM deviation meter with a variable frequency oscillator will also include the automatic frequency control voltage from the discriminator to allow

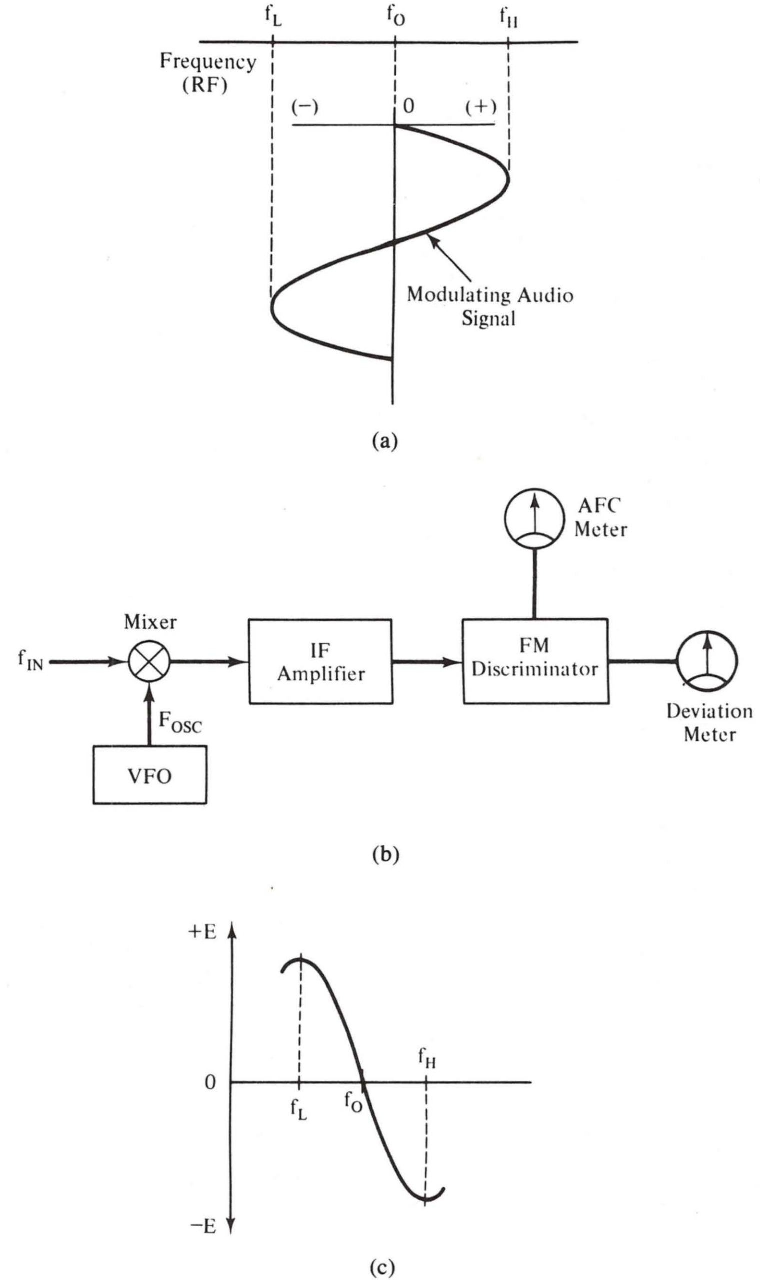


Figure 22-10. (a) Relative carrier frequency-vs-modulating audio amplitude. (b) FM deviation meter. (c) FM discriminator curve.

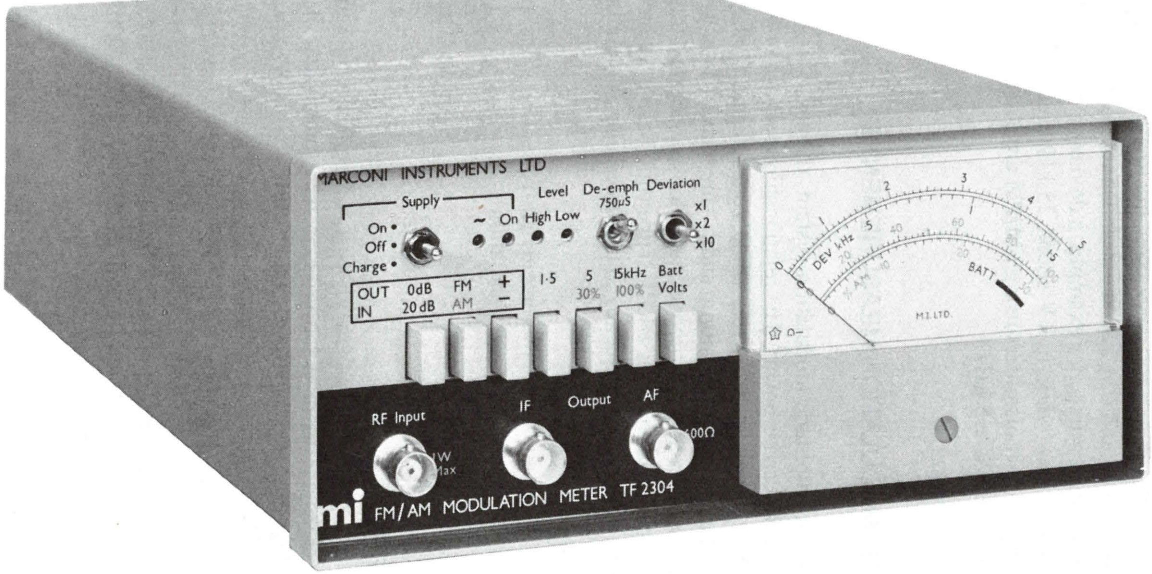


Figure 22-10. *(cont.)* (d) FM/AM modulation meter. (Courtesy of Marconi Instruments)

easier adjustment of the meter. The VFO is adjusted, with the transmitter turned on, for a zero output indication on the *AFC* meter. The transmitter is then modulated, and the deviation measured.

The term "100 percent modulation" has physical meaning in AM systems, but in FM and PM systems it is defined by regulation or general agreement. In FM broadcasting, for example, a deviation of ±75 kHz is defined as 100 percent modulation, while in TV sound 25 kHz qualifies as "100 percent." Most two-way radio transmitters using FM/PM are considered 100 percent modulated if the deviation is ±5 kHz.

22-10 AM MODULATION MEASUREMENTS

In an amplitude modulation system, the audio signal adds to, or subtracts from, the RF carrier power. The relationship is shown in Fig. 22-11. Zone A represents the unmodulated carrier, while zones B and C show the maxima and minima created by a modulating audio sine wave. This modulation envelope can be viewed on an oscilloscope that is loosely coupled to the RF by a "gimmick," or single-turn link. The oscilloscope time base is adjusted to show one or cycles of the modulating frequency. The percentage of modulation is given by:

$$\text{Percentage of modulation} = \frac{C - B}{C + B} \times 100\% \qquad (22\text{-}17)$$

Example 22-3

An oscilloscope with a 10 × 10 cm screen graticule is used to measure the percentage of modulation of an AM transmitter. It is found that dimension C [Fig. 22-11(a)] is 8.6 cm, and dimension B is 2.4 cm. Calculate the percentage of modulation.

SOLUTION

$$\text{Percentage of modulation} = (C - B)(100\%)/(C + B) \qquad (22\text{-}17)$$
$$\text{Percentage of modulation} = (8.6 - 2.4)(100\%)/(8.6 + 2.4)$$
$$\text{Percentage of modulation} = (6.2)(100\%)/(11) = \mathbf{56\%}$$

An alternate form of Eq. (22-17) is also used to measure the percentage of modulation. This other procedure is essentially the same, but uses somewhat different dimensions (see Sec. 18-12).

Another method for measuring amplitude modulation is shown in Fig. 22-12(a). The RF envelope is applied to the vertical input of the oscilloscope, and a

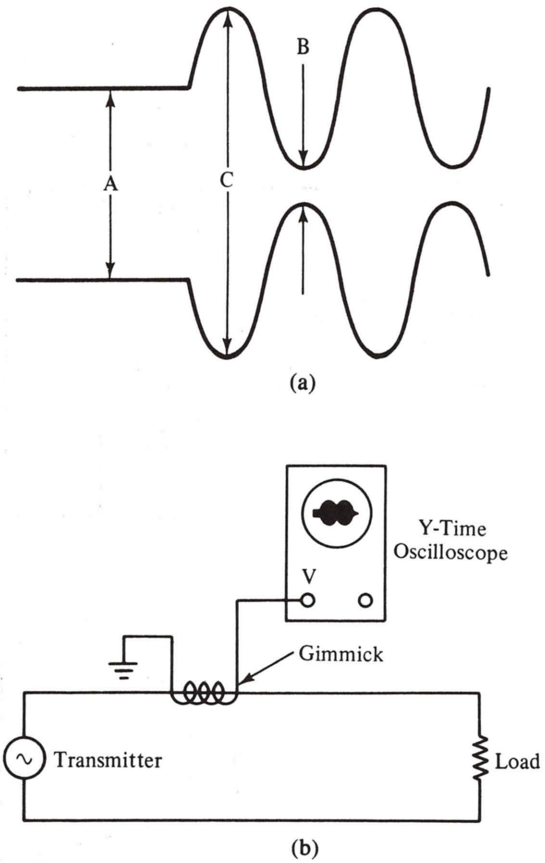


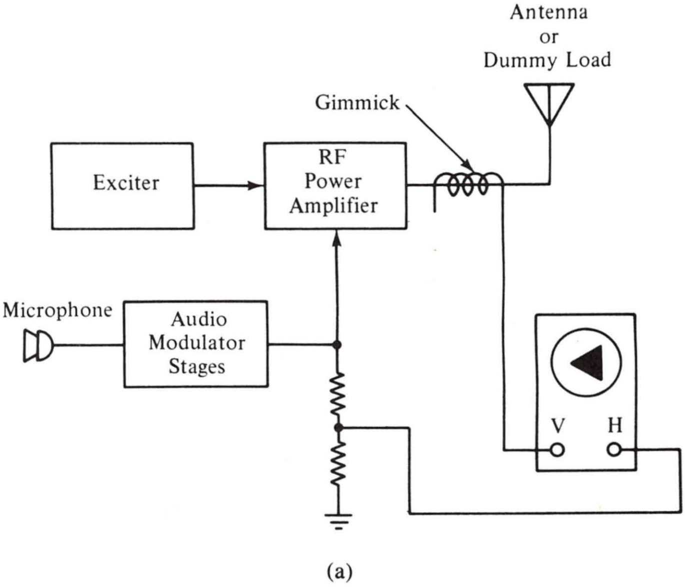
Figure 22-11. (a) AM waveform. (b) Applying AM waveform to Y-time oscilloscope.

sample of the modulating audio is applied to the horizontal input. (An X-Y oscilloscope is needed.) The trapezoidal pattern of Fig. 22-12(b) results when a sinusoidal audio signal is applied to the modulator. Using the notation of Fig. 22-12(b), we may compute the percentage of modulation from

$$\text{Percentage of modulation} = \frac{A - B}{A + B} \times 100\% \tag{22-18}$$

Example 22-4

Calculate the percentage of modulation if dimension A occupies 4.6 divisions on the oscilloscope screen and dimension B occupies 1.4 divisions.



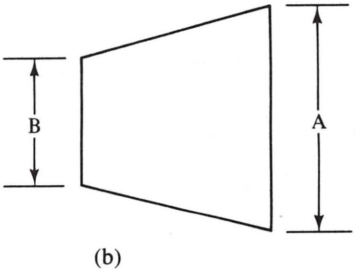


Figure 22-12. (a) Equipment connection for measuring AM modulation percentage by the trapezoid method. (b) AM trapezoidal trace.

SOLUTION

$$\text{Percentage of modulation} = (A - B)(100\%)/(A + B) \qquad (22\text{-}18)$$
$$\text{Percentage of modulation} = (4.6 - 1.4)(100\%)/(4.6 + 1.4)$$
$$\text{Percentage of modulation} = (3.2)(100\%)/(6) = \mathbf{53\%}$$

Figure 22-13 shows trapezoidal patterns for less than 100 percent, 100 percent, and greater than 100 percent modulation.

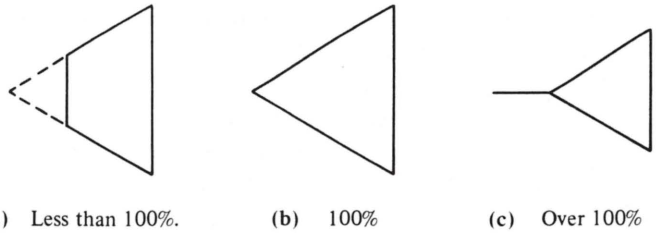


Figure 22-13. Trapezoidal traces for (a) less than 100 percent modulation, (b) 100 percent modulation, (c) over 100 percent modulation.

22-11 SSB MEASUREMENTS

A single sideband suppressed carrier (SSBSC, or simply, SSB) transmitter is a special case of the AM transmitter. Figure 22-14(a) shows the frequency spectrum for a regular AM transmitter, in this case a 1000-kHz carrier modulated by a 1-kHz audio sine wave.

AM is a multiplication process in which the audio and the RF carrier signals are heterodyned together, resulting in the production of sum and difference frequencies. The sum and difference are RF frequencies, and are called *sidebands*. In Fig. 22-14(a), we see the 1000-kHz carrier (F_0), the sum frequency F_2 of 1001 kHz, and a difference frequency F_1 of 999 kHz.

In SSB transmitters, carrier frequency F_0 and *one* of the two sidebands are suppressed, so only one sideband is transmitted. The spectrum of an upper

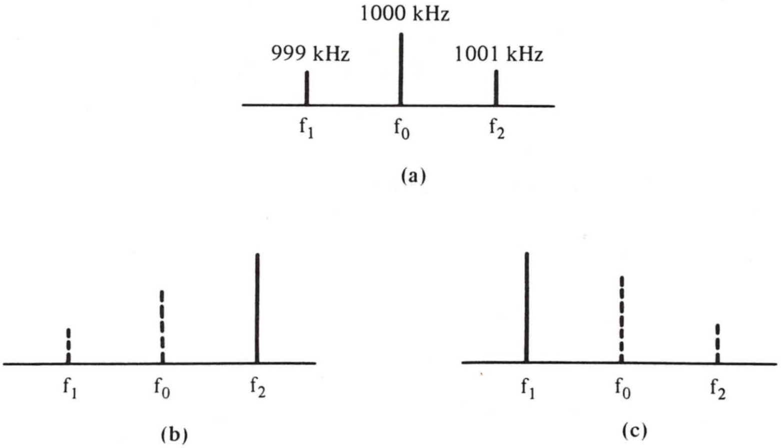


Figure 22-14. (a) AM signal. (b) USB signal. (c) LSB signal.

sideband (USB) transmitter is shown in Fig. 22-14(b). F_0 and F_1 have been suppressed, and only F_2 is transmitted.

Similarly, the spectrum for a lower sideband (LSB) transmitter is shown in Fig. 22-14(c). In this case, F_0 and F_2 are suppressed and only F_1 is transmitted.

All stages of an SSB transmitter past the modulator must be linear amplifiers. As a result of this requirement, it is necessary to use a two-tone test signal when one is making SSB measurements. The two tones are used to modulate the SSB transmitter, and the output signal is inspected for the existence of intermodulation distortion products, or flattening of the waveform. Figure 22-15(a) shows a normal two-tone output wave, using a transmitter connected to an oscilloscope as in Fig. 22-11(b). Note that the SSB percentage of modulation has no real meaning, as it does in straight AM, and the transmitter audio control is turned up until the output signal begins "flat-topping" [Fig. 22-15(b)]. The gain is then backed off until the flattening disappears.

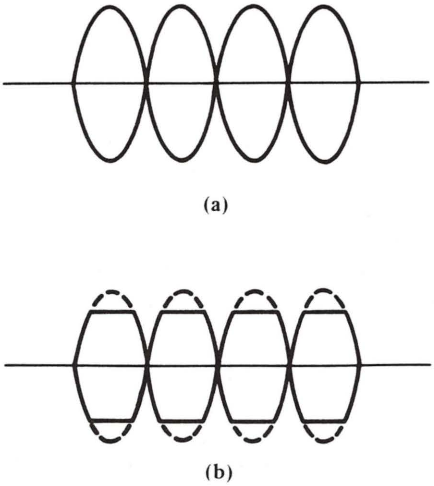


Figure 22-15. (a) Normal. (b) Clipped.

A spectrum analyzer will show intermodulation distortion by the creation of tone products other than the two tones applied to the transmitter audio input.

Connection of the transmitter is similar to Fig. 22-12(a), in which the RF amplifier output is fed to the vertical input of the oscilloscope, and the audio signal is applied to the horizontal input of the oscilloscope. The pattern produced by a two-tone test signal will be the "bow-tie" of Fig. 22-16. Distortion will show up as bending of the ordinarily straight and sharp edges of the pattern. This technique is also used to test linear amplifiers.

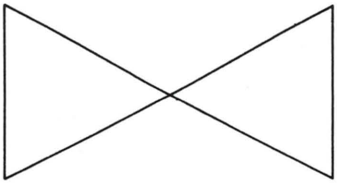


Figure 22-16. Two-tone "bow tie" trapezoidal pattern.

Figure 22-17(a) shows another test configuration for checking the linearity of SSB power amplifiers. This technique uses Lissajous figures on an oscilloscope. The oscilloscope gains are adjusted for approximately equal deflection amplitudes on both vertical and horizontal channels. A perfectly linear RF amplifier produces the pattern of Fig. 22-17(b), whereas nonlinearity shows up as a bending of the line, as in Fig. 22-17(c). Phase shift will show up as a thickening of the line, or as an oval trace, depending upon the severity.

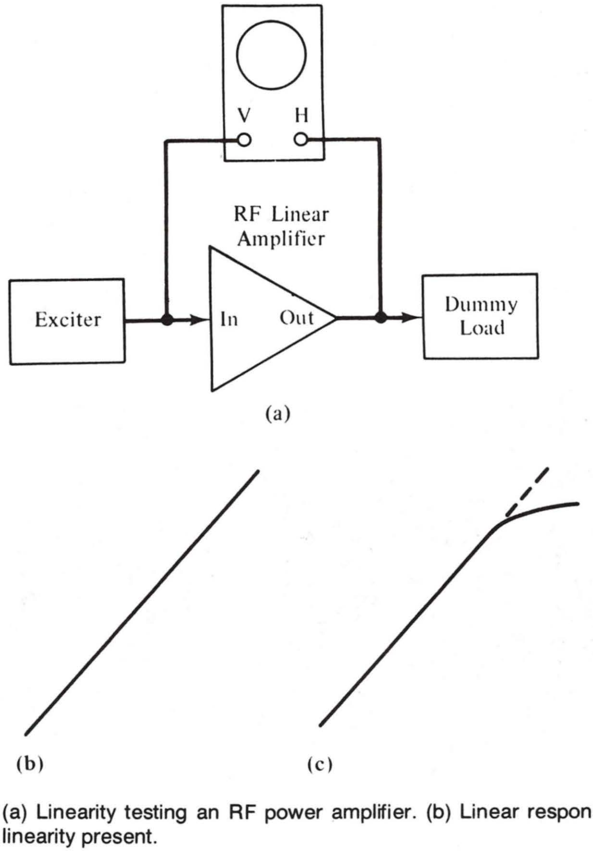


Figure 22-17. (a) Linearity testing an RF power amplifier. (b) Linear response. (c) Nonlinearity present.

22-12 COMMUNICATIONS MONITORS

Figure 22-18(a) shows the Lampkin model 107C communications monitor. This instrument, used mostly by two-way radio shops, can measure frequency from 10 kHz to over 1000 mHz, and serves as a signal generator between 1 kHz and 1000 mHz. The accuracy of the measurements will be on the order of 0.5 ppm, so will surpass the minimum FCC requirements for transmitter certification. The instrument also provides modulation for making receiver measurements, and produces CW (i.e., unmodulated), AM, and FM outputs. The power supply is capable of operating from 115 V ac mains power, or any of several dc voltages commonly found in landmobile, marine, or aviation systems.

The model 107C is a heterodyne frequency meter, and has a very useful *parts per million* dial. The transmitter's assigned frequency is set by using the three main frequency control knobs. The PPM dial is then adjusted for zero-beat with the transmitter's actual frequency. The transmitter frequency error in PPM can be read directly from the PPM dial, eliminating the need for computing the error from the actual frequency.

Two monitors that use the phase-locked loop frequency synthesizer are shown in Figs. 22-18(b) and 22-18(c). The Cushman model CE-5 is shown in Fig. 22-18(b), while a Motorola model is shown in Fig. 22-18(c). In both cases, frequency is selected by switches on the front panel. In these models, the Cushman is set to 125.000 mHz, while the Motorola is set to 164.5550 mHz. In both models, a *kilohertz error* meter and FM deviation meter are provided, along with an oscilloscope to view the modulation deviation characteristics of the transmitter. These instruments combine deviation and frequency measurement capability in one cabinet. Figure 22-18(d) shows a similar instrument, also by Motorola, in which the frequency meter section is a digital counter.

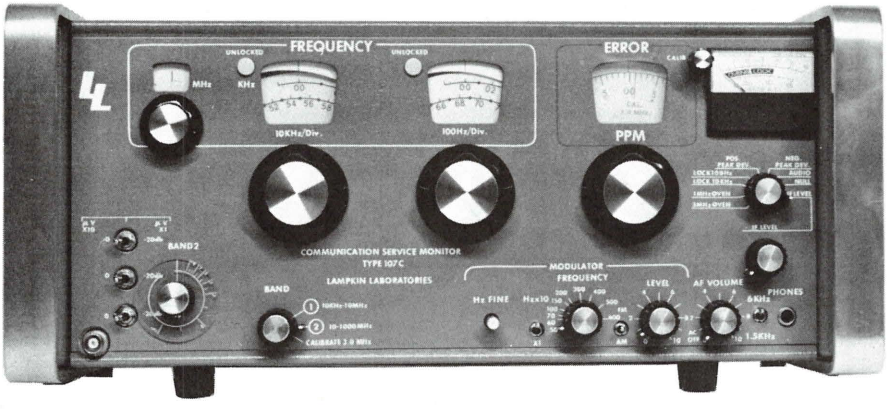
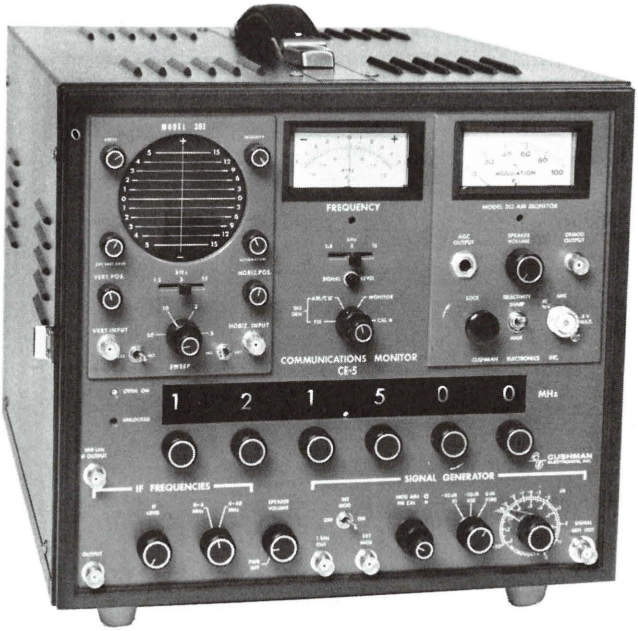
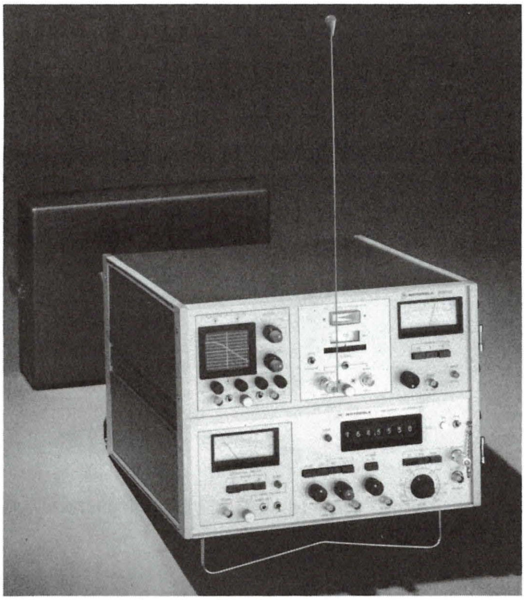


Figure 22-18. (a) Motorola Communications Monitor. (Courtesy of Motorola Communications.)



(b)



(c)

Figure 22-18. *(cont.)* (b) (c) Motorola Communications Monitor. (Courtesy of Motorola Communications.)

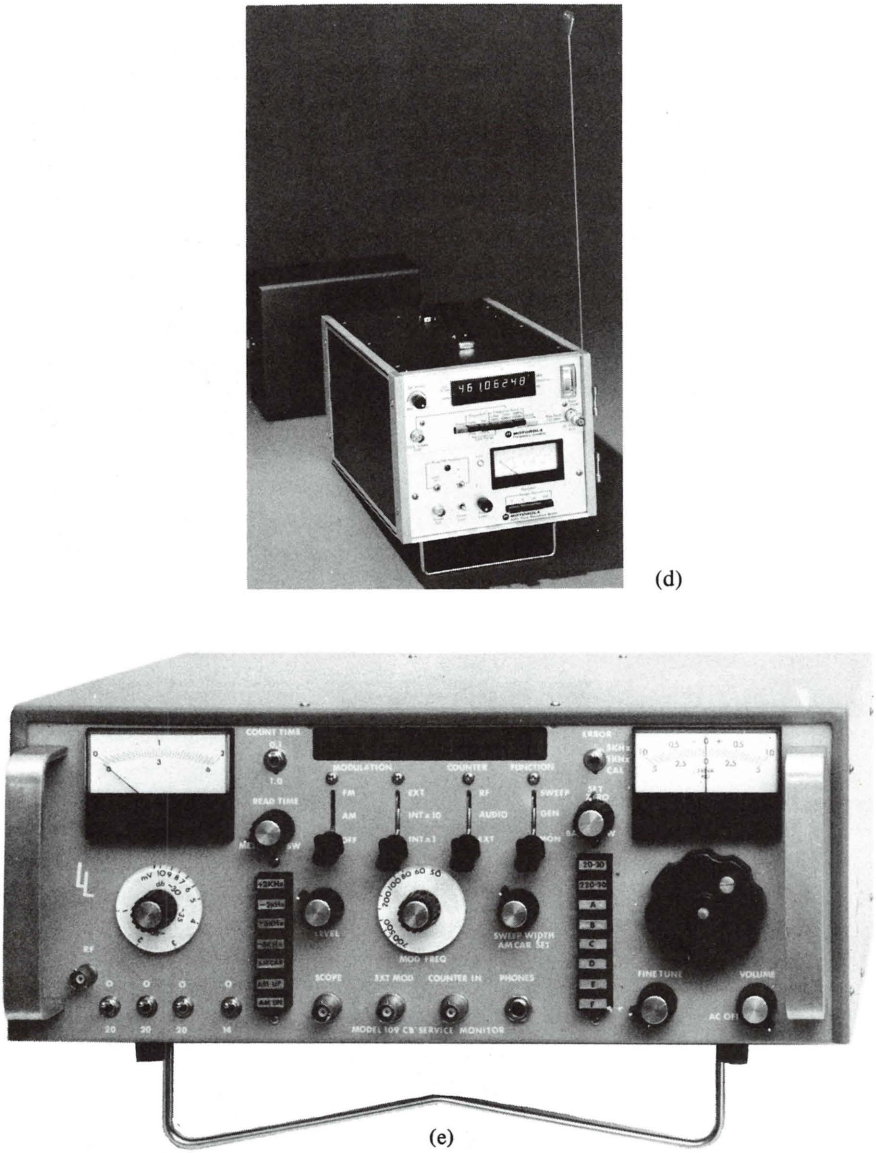


Figure 22-18. *(cont.)* (d) (e) CB Communications Monitor. (Courtesy of Hampkin Laboratories, Inc.)

A Lampkin model 109 CB service monitor is shown in Fig. 22-18(e). This type of monitor has become popular in recent years because of the tremendous expansion in CB activity. The instrument contains a digital frequency counter, and a signal generator covering commonly used frequencies (i.e., RF, IF, local oscillator, etc.).

22-13 SUMMARY

1. Simple voltage-bridge RF wattmeters produce good results if CW or FM signals are applied, and if you have sinusoidal waveshape.
2. Thermal RF wattmeters and thermocouple RF ammeters produce true-rms readings.
3. A dummy load is a shielded, nonreactive resistance with a value equal to the expected load impedance of the transmitter, or surge impedance of the transmission line.
4. FM deviation meters are essentially FM receivers with a discriminator output voltmeter calibrated in units of frequency.
5. AM modulation may be checked on an oscilloscope by viewing either the trapezoidal pattern or a modulation envelope display.

22-14 RECAPITULATION

Now go back and try answering the questions at the beginning of the chapter. When you are finished, answer the questions and work the problems given below. Place a mark beside each problem or question that you cannot answer, and then go back to the text and reread appropriate sections.

22-15 QUESTIONS

1. How is RF power measured with an ammeter?
2. List two types of RF power measurement that yield an rms result.
3. Draw a schematic for two types of simple RF wattmeter. State any limitations of these instruments.
4. Describe what is meant by "dummy load" in your own words.
5. An instrument that mates a dummy load with an RF wattmeter is called an ____________.
6. How is percentage of modulation defined for (a) AM and (b) FM?
7. In SSB transmitters a ____________ ____________ audio signal is used for modulation measurements.
8. A principal measurement of SSB transmitter linearity is the ____________ distortion present in the output waveform.
9. A ____________ Lissajous pattern gives an indication of SSB distortion.

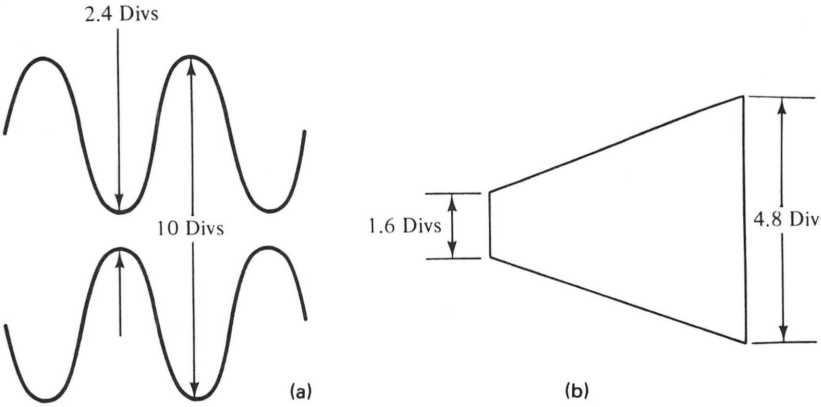


Figure 22-19.

22-16 PROBLEMS

1. A transmitter causes 2.85 amperes of RF current to flow in a 36-ohm resistive load. Calculate the RF power level.
2. Calculate the rms power of a sine wave that has a peak potential of 52 volts across a 50-ohm load.
3. A transmitter produces a reading of 98 mW in a thermal power meter when 22 dB of attenuation is used. Calculate the transmitter output power.
4. Calculate the percentage of modulation indicated on the RF envelope in Fig. 22-19(a).
5. Calculate the percentage of modulation indicated by the trapezoidal pattern of Fig. 22-19(b).

chapter 23

GENERAL PURPOSE INTERFACE BUS INSTRUMENTS

23-1 OBJECTIVES

1. To learn the details of the IEEE-488 instruments.
2. To learn how to use IEEE-488 bus instruments.

23-2 SELF-EVALUATION

Before studying the material in this chapter, try answering the questions given below. These questions test your knowledge of the subject. If you cannot answer a particular question, then look for the answer as you read the text.

1. The general purpose interface bus is used primarily in _____ _____ equipment applications.
2. Name the three major buses in the GPIB: _____, _____, _____;
3. Name the two basic configurations for an IEEE-488 instrumentation system.
4. What is the function of the DI01 through DI08 lines?

23-3 INTRODUCTION TO THE GPIB

Automatic test equipment is now one of the leading methods for testing electronic equipment in factory production and troubleshooting situations. The basic method is to use a programmable digital computer to control a bank of test instruments. The program, often in the BASIC language, turns the various instruments on and off and then evaluates the results as measured by other instruments.

The bank of equipment (see Fig. 23-1 for an example) can be configured for a special purpose or for general use. For example, we could select a particular line-up of equipment needed to test, say, a broadcast audio console and provide a computer program to make the various measurements: gain, frequency response,

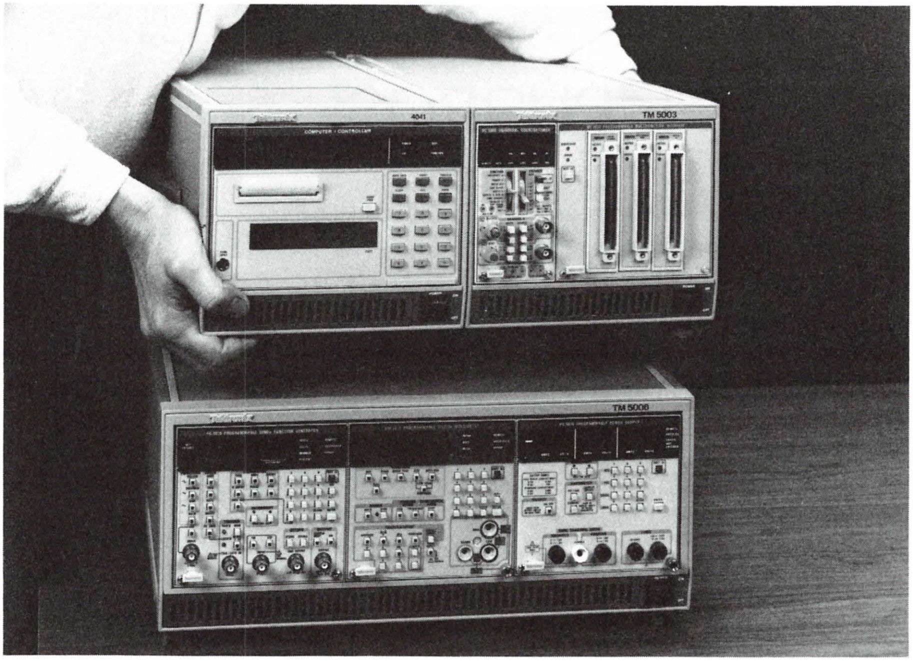


Figure 23-1. Automatic test equipment system using GPIB. (Courtesy of Tektronix, Inc.)

total harmonic distortion, and so forth. Alternatively, we could also make a generalized test set. This is the method selected by a number of organizations who have large amounts of different electronic devices to test. There will be a main bank of electronic test equipment, adapters to make the devices under test interconnect with the system, and a special program for each type of equipment. Such an approach makes for a cost-effective system of test equipment.

Previously, the main problem in attempting to make automatic test equipment (ATE) was that it was impossible to use unmodified off-the-shelf commercial test instruments. Thus, many of the best and most useful pieces of test equipment could not be used at all, or had to be extensively modified by the ATE maker before they could be used. The problem was programmability—how to make a signal generator respond to the computer commands. In 1978, the Institute of Electrical and Electronic Engineers (IEEE) released their specification titled *IEEE Standard Digital Interface for Programmable Instrumentation*, or IEEE-488 as it is called in the trade. This specification provides details for a standard computer interface between a computer and instruments. It also calls out ASCII codes and mnemonics for program instructions. The IEEE-488 bus is also called a *general purpose interface bus* (GPIB). The Hewlett-Packard interface bus (HPIB) is a proprietary version of the IEEE-488 bus. The main purpose for the IEEE-488/GPIB is automatic test equipment, both generalized and specific.

Test instrumentation that is intended for GPIB service will have a 24-pin "blueline" connector on the rear panel. This connector is one of the Amphenol blueline series not unlike the 36-pin connector used for parallel printer interface on microcomputers. There will also be a GPIB ADDRESS DIP switch on the rear panel, usually near the connector. The purpose of the switch is to set the 5-bit binary address where the instrument is located in the system; it determines whether or not the device is a listener only or a talker only and certain other details.

23-4 GPIB BASICS

The IEEE-488/GPIB specification provides technical details of the standard bus. The logic levels on the bus are generally similar to TTL: a LOW is less than or equal to 0.800 volts, while a HIGH is greater than 2.0 volts. The logic signals can be connected to the instruments through a multiconductor cable up to 20 meters (66 feet) in length, provided that an instrument load is placed every 2 meters. This specification works out to a cable length in meters twice the number of instruments in the system. Most IEEE-488/GPIB systems operate unrestricted to 250 kilobytes per second or faster with certain specified restrictions.

There are two basic configurations for the IEEE-GPIB system (Fig. 23-2): linear and star. These configurations are created with the cable connections

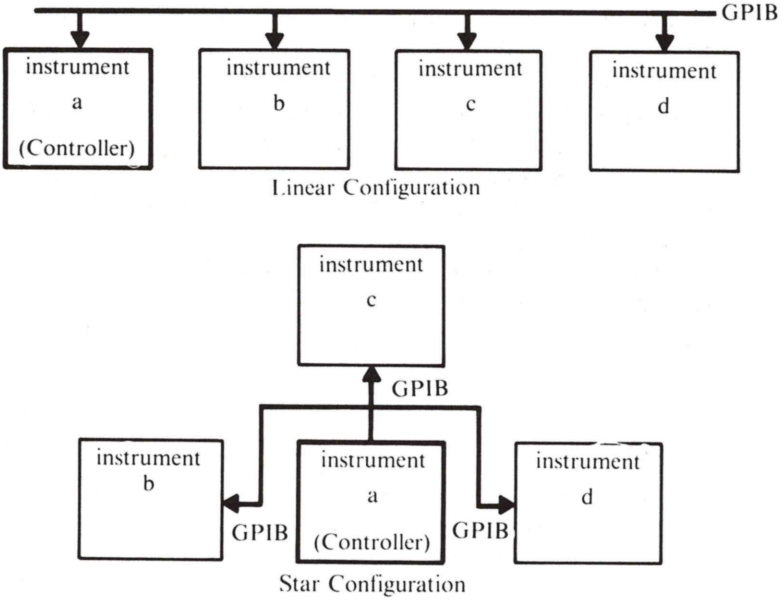


Figure 23-2. GPIB system configurations. (Courtesy of Tektronix, Inc.)

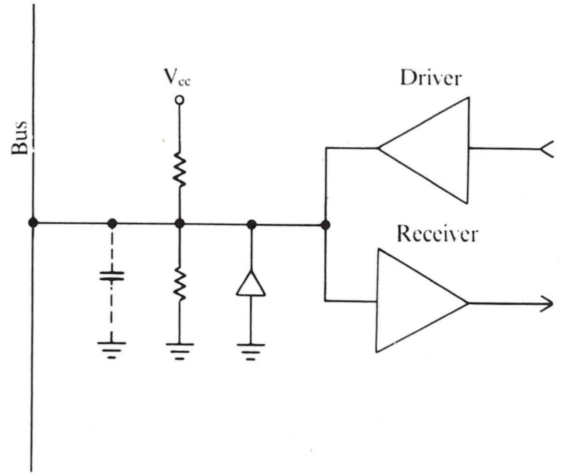


Figure 23-3. IEEE-488/GPIB input/output circuits.

between the instruments and the computer. The linear configuration is basically a daisy chain method, in which the tap-off to the next instrument is taken from the previous one in the series. In the star configuration, the instruments are connected from a central point.

There are three major buses in the IEEE-488/GP1B system. Each line in each bus has a circuit similar to that shown in Fig. 23-3. Besides the shunt protection diode and stray capacitance, there are also pull-up and pull-down resistors that effectively determine the standardized input impedance. Connected to the bus line are receiver and driver circuits. These similar-to-TTL logic elements provide input or output to the instrument. The driver is an output and will be a tristate device. That is, it is inert until commanded to turn on. A tristate output will float at high impedance until turned on. The receiver is basically a noninverting buffer with a high impedance input. This arrangement of drivers and receivers provides low loading to the bus.

Figure 23-4 shows the basic structure of the IEEE-488/GPIB. There are three buses and four different types of devices. The devices are controllers, talker only, listener only, and talker/listener. These devices are defined as follows:

Controllers. This type of device acts as the brain of the system and communicates device addresses and other interface messages to instruments in the system. Most controllers are programmable digital computers. Both Hewlett-Packard and Tektronix offer computers that serve this function, and certain other companies produce hardware and software that permit other computers to act as IEEE-488/GPIB controllers.

Listener. A device capable of listening will receive commands from another instrument, usually the controller, when the correct address is placed on the

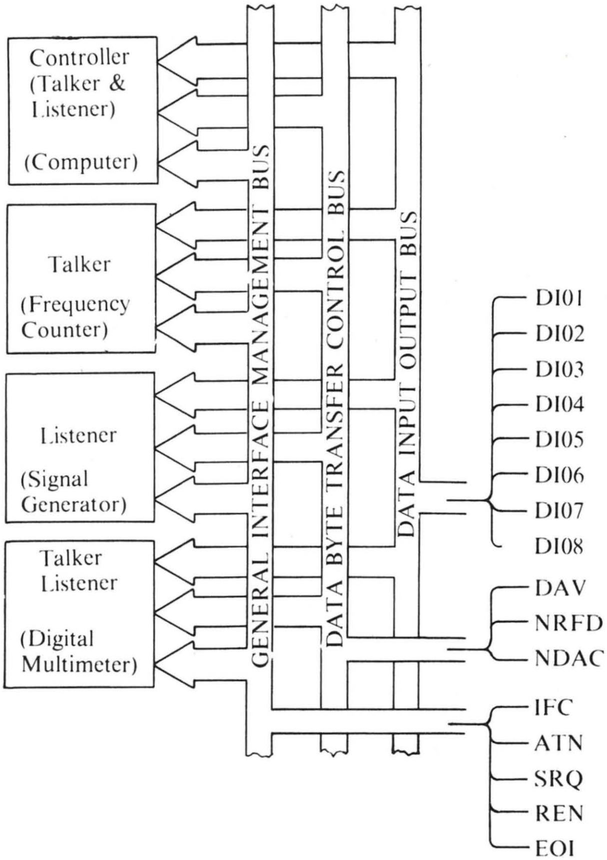


Figure 23-4. IEEE-488/GPIB structure.

bus. The listener acts on the message received, but does not send back any data to the controller. As shown in Fig. 23-4, is the signal generator is an example of a listener.

Talker. The talker responds to the message sent to it by the controller and then sends data back to the controller over the DIO data bus. A frequency counter is an example of a talker.

There is also a combination device that accepts commands from the controller to set up ranges, tasks, etc., and then returns data back over the DIO bus to the controller. An example is a digital multimeter (DMM). The controller will send the DMM commands that determine whether it is ac or dc, volts or milliamperes or ohms, and what specific range—the device is thus acting as a listener. When the measurement is made, the DMM becomes a talker and transmits the data measured back over the DIO bus to the controller.

There are three major buses in the IEEE-488/GPIB system: general interface management (GIM), data byte transfer (DBT), and the data input/output (DIO) Bus. These buses operate as described below.

DIO bus. The data input/output bus is a bidirectional 8-bit data bus that carries data, interface messages, and device-dependent messages between the controller, talkers, and listeners. This bus sends data asynchronously in byte-serial format.

DBT bus. The data byte transfer bus controls the sending of data along the DIO bus. There are three lines in the DBT bus: data valid (DAV), not ready for data (NRFD), and not data accepted (NDAC). These signal lines are defined as:

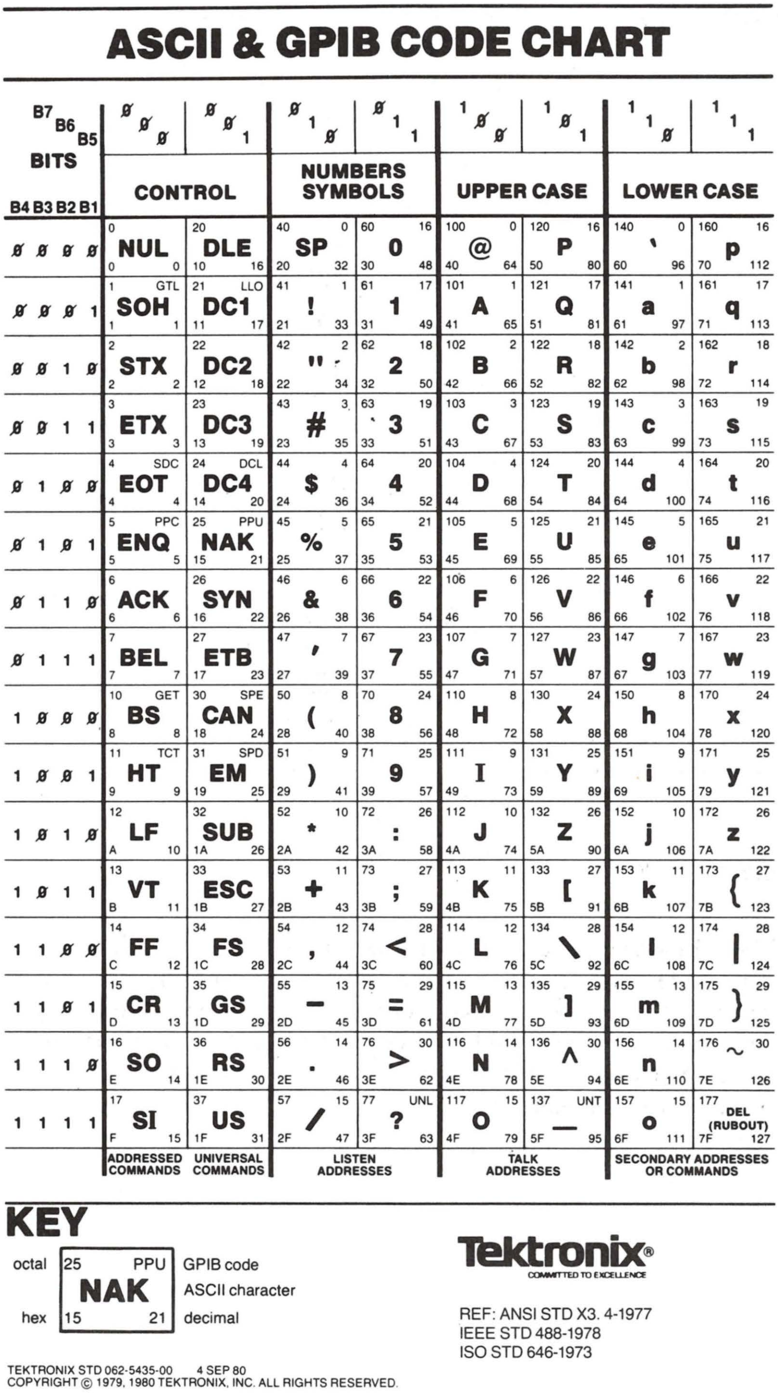
DAV. The data valid signal indicates the availability and validity of the data on the line. If the measurement is not finished, for example, the DAV signal will be false.
NRFD. The not ready for data signal lets the controller know whether or not the specific device addressed is in a condition to receive data.
NDAC. The not data accepted signal line is used to indicate to the controller whether or not the device accepted the data sent to it over DIO bus.

GIM bus. The general interface management bus coordinates the system and ensures an orderly flow of data over the DIO bus; it has the following signals: interface clear (IFC), attention (ATN), service request (SRQ), remote enable (REN), and end or identify (EOI). These signals are defined as follows:

IFC. The interface clear signal is used by the controller to place all devices in a predefined quiescent or standby condition.
ATN. The attention signal is used by the controller/computer to let the system know how data on the DIO bus lines is to be interpreted, and which device is to respond to the data.
SRQ. The service request signal is used by a device on the system to ask the controller for attention. This signal is essentially an interrupt request.
REN. The remote enable signal is used by the controller to select from between two alternate sources of device programming data.
EOI. The end or identify signal is used by talkers for two purposes. It will follow the end of a multiple byte sequence of data in order to indicate that the data is now finished. It is also used in conjunction with the ATN signal for polling the system.

The 7-bit binary signals used in the IEEE-488/GPIB system for ASCII and GPIB message codes are shown in Fig. 23-5.

The signals defined above are implemented as conductors in a system interface cable. Each IEEE-488/GPIB compatible instrument will have a female 36-pin Amphenol-style connector on the rear panel. The pinout definitions are as follows:



ASCII & GPIB CODE CHART

B7 B6 B5	0 0 0	0 0 1	0 1 0	0 1 1	1 0 0	1 0 1	1 1 0	1 1 1
BITS B4 B3 B2 B1	CONTROL		NUMBERS SYMBOLS		UPPER CASE		LOWER CASE	
0 0 0 0	0 NUL 0 0	20 DLE 10 16	40 0 SP 20 32	60 16 0 30 48	100 0 @ 40 64	120 16 P 50 80	140 0 ` 60 96	160 16 p 70 112
0 0 0 1	1 GTL SOH 1 1	21 LLO DC1 11 17	41 1 ! 21 33	61 17 1 31 49	101 1 A 41 65	121 17 Q 51 81	141 1 a 61 97	161 17 q 71 113
0 0 1 0	2 STX 2 2	22 DC2 12 18	42 2 " 22 34	62 18 2 32 50	102 2 B 42 66	122 18 R 52 82	142 2 b 62 98	162 18 r 72 114
0 0 1 1	3 ETX 3 3	23 DC3 13 19	43 3 # 23 35	63 19 3 33 51	103 3 C 43 67	123 19 S 53 83	143 3 c 63 99	163 19 s 73 115
0 1 0 0	4 SDC EOT 4 4	24 DCL DC4 14 20	44 4 $ 24 36	64 20 4 34 52	104 4 D 44 68	124 20 T 54 84	144 4 d 64 100	164 20 t 74 116
0 1 0 1	5 PPC ENQ 5 5	25 PPU NAK 15 21	45 5 % 25 37	65 21 5 35 53	105 5 E 45 69	125 21 U 55 85	145 5 e 65 101	165 21 u 75 117
0 1 1 0	6 ACK 6 6	26 SYN 16 22	46 6 & 26 38	66 22 6 36 54	106 6 F 46 70	126 22 V 56 86	146 6 f 66 102	166 22 v 76 118
0 1 1 1	7 BEL 7 7	27 ETB 17 23	47 7 ' 27 39	67 23 7 37 55	107 7 G 47 71	127 23 W 57 87	147 7 g 67 103	167 23 w 77 119
1 0 0 0	10 GET BS 8 8	30 SPE CAN 18 24	50 8 (28 40	70 24 8 38 56	110 8 H 48 72	130 24 X 58 88	150 8 h 68 104	170 24 x 78 120
1 0 0 1	11 TCT HT 9 9	31 SPD EM 19 25	51 9) 29 41	71 25 9 39 57	111 9 I 49 73	131 25 Y 59 89	151 9 i 69 105	171 25 y 79 121
1 0 1 0	12 LF A 10	32 SUB 1A 26	52 10 * 2A 42	72 26 : 3A 58	112 10 J 4A 74	132 26 Z 5A 90	152 10 j 6A 106	172 26 z 7A 122
1 0 1 1	13 VT B 11	33 ESC 1B 27	53 11 + 2B 43	73 27 ; 3B 59	113 11 K 4B 75	133 27 [5B 91	153 11 k 6B 107	173 27 { 7B 123
1 1 0 0	14 FF C 12	34 FS 1C 28	54 12 , 2C 44	74 28 < 3C 60	114 12 L 4C 76	134 28 \ 5C 92	154 12 l 6C 108	174 28 | 7C 124
1 1 0 1	15 CR D 13	35 GS 1D 29	55 13 − 2D 45	75 29 = 3D 61	115 13 M 4D 77	135 29] 5D 93	155 13 m 6D 109	175 29 } 7D 125
1 1 1 0	16 SO E 14	36 RS 1E 30	56 14 . 2E 46	76 30 > 3E 62	116 14 N 4E 78	136 30 ^ 5E 94	156 14 n 6E 110	176 30 ~ 7E 126
1 1 1 1	17 SI F 15	37 US 1F 31	57 15 / 2F 47	77 UNL ? 3F 63	117 15 O 4F 79	137 UNT _ 5F 95	157 15 o 6F 111	177 DEL (RUBOUT) 7F 127
	ADDRESSED COMMANDS	UNIVERSAL COMMANDS	LISTEN ADDRESSES		TALK ADDRESSES		SECONDARY ADDRESSES OR COMMANDS	

KEY

octal	25 PPU	GPIB code
	NAK	ASCII character
hex	15 21	decimal

COMMITTED TO EXCELLENCE

REF: ANSI STD X3. 4-1977
IEEE STD 488-1978
ISO STD 646-1973

TEKTRONIX STD 062-5435-00 4 SEP 80

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Figure 23-5. ASCII and GPIB codes (Courtesy of Tektronix, Inc.)

Pin No.	*Signal Line*
1	DIO1
2	DIO2
3	DIO3
4	DIO4
5	EOI
6	DAV
7	NRFD
8	NDAC
9	IFC
10	SRQ
11	ATN
12	shield
13	DIO5
14	DIO6
15	DIO7
16	DIO8
17	REN
18	ground (6)
19	ground (7)
20	ground (8)
21	ground (9)
22	ground (10)
23	ground (11)
24	logic ground

23-5 SUMMARY

The IEEE-488 general purpose interface bus (GPIB) can be used to marry together various pieces of standard commercial test equipment, computers, and customized equipment into a system that is capable of providing specified tests and measurements under the control of a computer program, often in BASIC. The GPIB makes either special purpose or generalized automatic test equipment (ATE) possible with a minimum of effort.

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